WIRELESS ENGINEER

THE JOURNAL OF RADIO RESEARCH & PROGRESS

SEPTEMBER 1954

VOL. 31 No. 9 THREE SHILLINGS AND SIXPENCE

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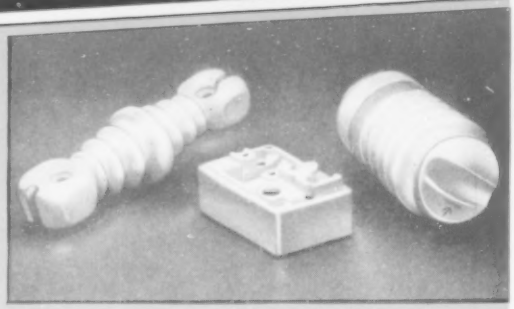

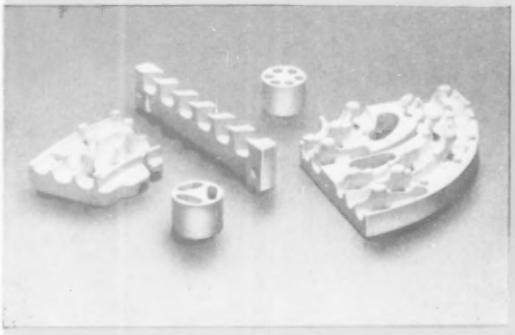

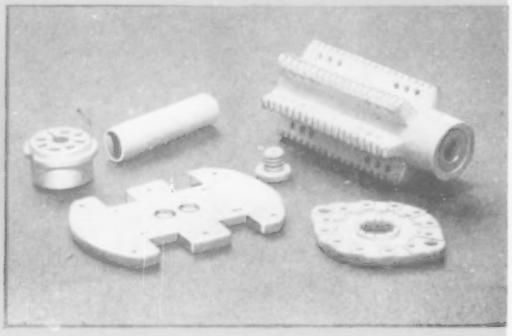

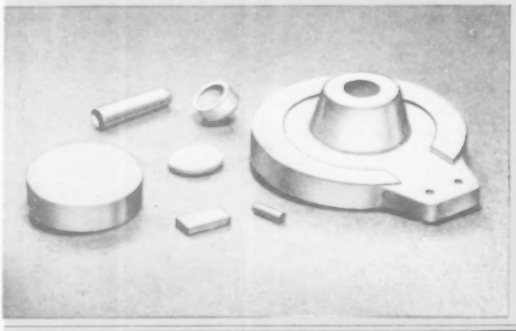

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S.G.2.

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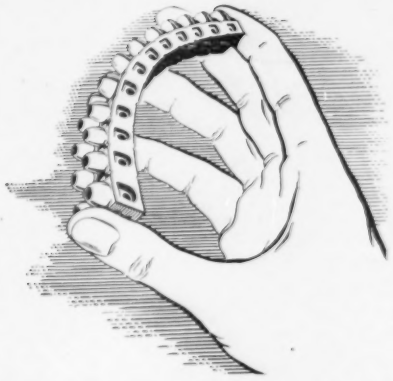
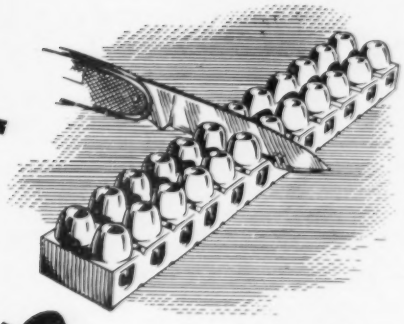


You can cut it..



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The component can be cut with an ordinary pocket knife into smaller groups of terminals as required, and for the purpose of fixing, holes are provided between each pair of terminals.

- ★ Captive Terminal Screws
- ★ Flexible
- ★ Easily Sectionalised
- ★ Mechanically Shock-proof
- ★ Rating 5 amp. Up to 10 amp may be used at designer's discretion
- ★ 12 Way
- ★ Overall size 5″ x .750″ x .625″ high. Holed centres spaced .425″

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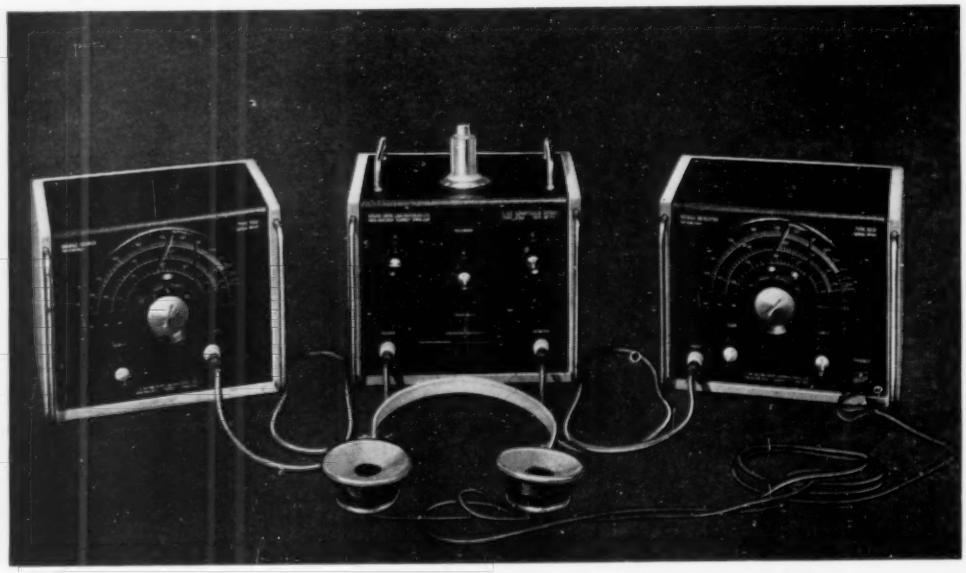

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G.E.C.

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BAND COVERAGE

0.150—0.385 Mc/s } in 6 bands
0.510—30.0 Mc/s }

SENSITIVITY

Better than 1.0 μV for 1.5 watts output, over the whole band.

SIGNAL/NOISE RATIO

Standard input for 20 db:—

1.3—30.0 Mc/s	...	...	... < 7.0 μV
0.150—1.3 Mc/s	...	...	... < 10.0 μV

SELECTIVITY

Six switched bandwidths:—

0.5 kc/s	1.0 kc/s	2.0 kc/s
5.5 kc/s	9.0 kc/s	13.0 kc/s

OVERALL FIDELITY

Less than 2 db down at 50 c/s
Less than 6 db down at 5,500 c/s

A.G.C. CHARACTERISTICS

Output constant within 3 db for 100 db change in signal input.

OUTPUT CIRCUITS

At 2.5 or 15 ohms	...	...	...	2.5 watts
At 120 ohms	...	...	...	0.05 watts
At 600 ohms	...	...	...	0.2 watts

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95—130 and 195—250 volts, 40/80 c/s. Also from 12 volt battery, using BRT 401 auxiliary power unit.

For the full specification please send for a copy of publication BC2084.

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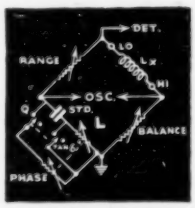

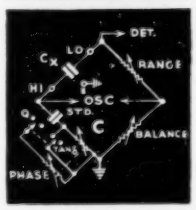

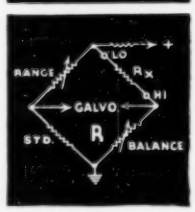

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Wings Day SEPT 18 — Give for those who Gave

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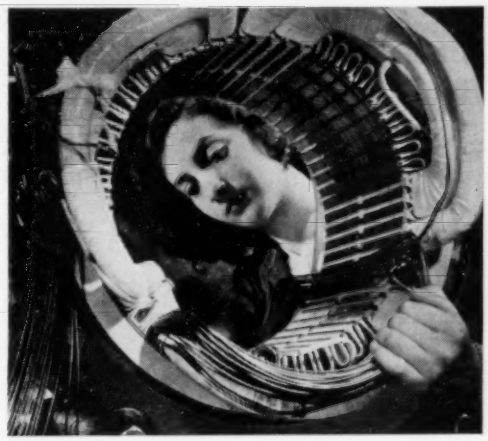

'Bicalex' Winding Wires give you all the well-recognized advantages of standard enamel winding wires, plus special advantages of their own. Some of these are listed on the right.

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- ★ Windings which are to be impregnated with quick-drying synthetic varnishes of high bonding strength.
- ★ Dry unimpregnated windings which may be subject to high humidity.

The dielectric strength of 'Bicalex' Winding Wires fully satisfies the requirements of B.S.1844. For fuller information write for Publication No. 322.

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THEY'RE TOUGH—to withstand rough handling

High-speed winding and its attendant rigours do not harm them. They can be stretched, twisted and even flattened without the insulation being damaged.

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THEY WITHSTAND 110°C WITHOUT "AGEING"

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THEIR COVERING WILL NOT "TUBE"

The covering of 'Bicalex' Wires adheres completely to the conductor, thus guarding against damage by "plucking", or "tubing" on elongation.

THEY RESIST CONTAMINATED AND HUMID ATMOSPHERES

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Photo by courtesy of Telegraph Condenser Company Ltd.

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epoxy casting resins

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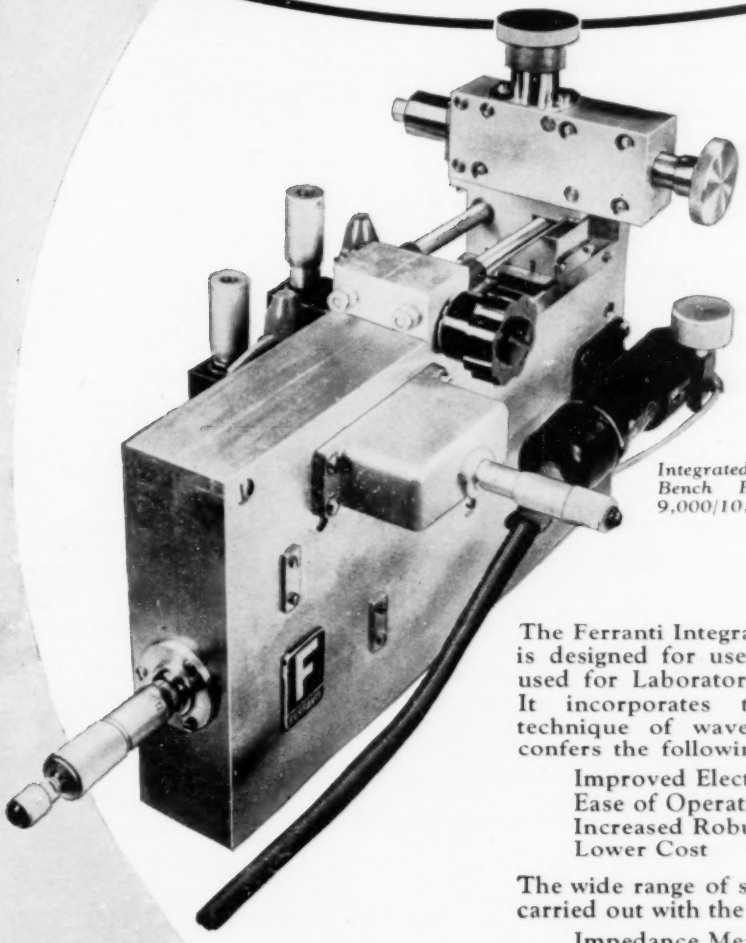

Integrated Microwave Test Bench Frequency Range 9,000/10,000 Mc/s.

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is the Sullivan-Griffiths Precision Inductance Bridge.

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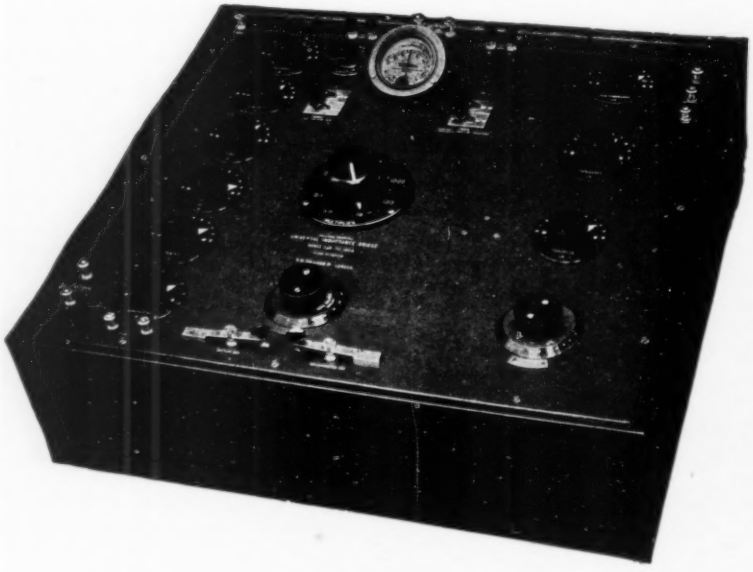

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	RR3—1250	CV2518	B4F	5.0	7.1	10	5.0	1.25	−55 to +70
MERCURY VAPOUR	RG1—240A	CV1626/1072	British 4-pin	4.0	2.7	6.5	1.25	0.25	+10 to +40
	RG3—250	CV1625	Medium Edison Screw	2.5	5.0	10	1.0	0.25	+10 to +40
	RG3—250A	CV32	4-pin UX	2.5	5.0	10	1.0	0.25	+10 to +40
	RG3—1250	CV1629/152	Goliath Edison Screw	4.0	7.0	13	5.0	1.25	+10 to +40
	RG4—1250	CV5	Goliath Edison Screw	4.0	11	13	5.0	1.25	+10 to +40
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WIRELESS ENGINEER

Vol. 31 SEPTEMBER 1954 No. 9

Rotating Observers

WE have on several occasions emphasized the absolute character of rotation, and we have just received from a continental professor a letter referring to the June Editorial and asking if we can give a reference to "something in the literature about the principles of your fundamental affirmation that rotation is absolute".

When two or more bodies are moving linearly at different constant speeds, and even in different directions, it is open to an observer on any one body to assume that he is at rest and that the other bodies are moving; any statement of the speed of a body has no meaning unless the frame of reference is stated or inferred. When discussing the induction of e.m.f. due to the relative motion of conductors and magnetic poles, so long as the motions are purely linear one can explain the phenomena as due either to the conductor moving while the pole is at rest, as it would appear to an observer situated on the pole, or *vice versa*, as it would appear to an observer on the conductor.

To an observer firmly stationed on a solid body every point on the body appears to be at rest even if it is rotating. If, to an observer on one body, two points on another body appear to be moving at different velocities, then this second body is rotating. If in Fig. 1 the point A is moving at the moment relatively to any observer in the direction shown with velocity v_a, while the point B on the line perpendicular to v_a appears to be moving in the same direction with velocity v_b, then the body is rotating about the point O, or, more simply, if the points A and C are moving at the moment in the directions shown, the intersection O of the perpendiculars gives the point about which the body is rotating at the moment and v_a/OA or v_c/OC gives the angular velocity of the rotation.

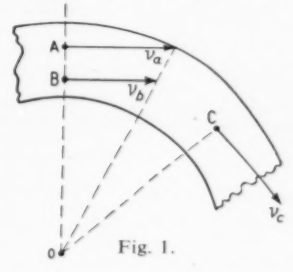

Fig. 1.

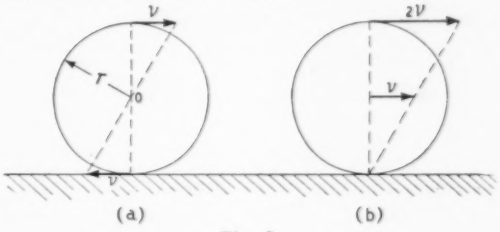

Fig. 2.

To take a well-known example, Fig. 2(a) shows the wheel of a vehicle as seen from the vehicle and Fig. 2(b) as seen by a stationary observer on the road. The rotation is the same in both cases, the angular velocity being v/r; to the observer on the vehicle this is the only motion, but to the stationary observer there is a superposed linear velocity v. This shows that although the linear velocity is relative, the angular velocity of rotation is unaffected by the movement of the observer. An observer on a rotating body cannot assume that he is at rest; as I have said before he will have feelings that belie such an assumption and, unless he is firmly secured to the body, when the speed of rotation reaches a certain value he will be flung off, whatever views he may hold on

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	RR3—1250	CV2518	B4F	5.0	7.1	10	5.0	1.25	— 55 to + 70
MERCURY VAPOUR	RG1—240A	CV1626/1072	British 4-pin	4.0	2.7	6.5	1.25	0.25	+ 10 to + 40
	RG3—250	CV1625	Medium Edison Screw	2.5	5.0	10	1.0	0.25	+ 10 to + 40
	RG3—250A	CV32	4-pin UX	2.5	5.0	10	1.0	0.25	+ 10 to + 40
	RG3—1250	CV1629/152	Goliath Edison Screw	4.0	7.0	13	5.0	1.25	+ 10 to + 40
	RG4—1250	CV5	Goliath Edison Screw	4.0	11	13	5.0	1.25	+ 10 to + 40
	* 872A	CV642	B4F	5.0	7.5	10	5.0	1.25	+ 10 to + 40

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WIRELESS ENGINEER

Vol. 31 SEPTEMBER 1954 No. 9

Rotating Observers

WE have on several occasions emphasized the absolute character of rotation, and we have just received from a continental professor a letter referring to the June Editorial and asking if we can give a reference to "something in the literature about the principles of your fundamental affirmation that rotation is absolute".

When two or more bodies are moving linearly at different constant speeds, and even in different directions, it is open to an observer on any one body to assume that he is at rest and that the other bodies are moving; any statement of the speed of a body has no meaning unless the frame of reference is stated or inferred. When discussing the induction of e.m.f. due to the relative motion of conductors and magnetic poles, so long as the motions are purely linear one can explain the phenomena as due either to the conductor moving while the pole is at rest, as it would appear to an observer situated on the pole, or *vice versa*, as it would appear to an observer on the conductor.

To an observer firmly stationed on a solid body every point on the body appears to be at rest even if it is rotating. If, to an observer on one body, two points on another body appear to be moving at different velocities, then this second body is rotating. If in Fig. 1 the point A is moving at the moment relatively to any observer in the direction shown with velocity v_a, while the point B on the line perpendicular to v_a appears to be moving in the same direction with velocity v_b, then the body is rotating about the point O, or, more simply, if the points A and C are moving at the moment in the directions shown, the intersection O of the perpendiculars gives the point about which the body is rotating at the moment and v_a/OA or v_c/OC gives the angular velocity of the rotation.

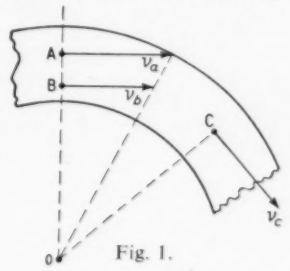

Fig. 1.

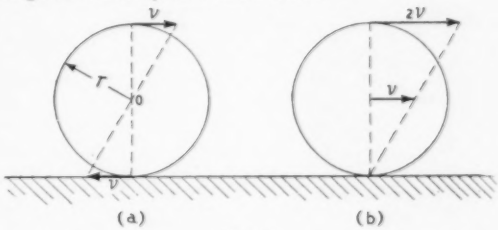

Fig. 2.

To take a well-known example, Fig. 2(a) shows the wheel of a vehicle as seen from the vehicle and Fig. 2(b) as seen by a stationary observer on the road. The rotation is the same in both cases, the angular velocity being v/r; to the observer on the vehicle this is the only motion, but to the stationary observer there is a superposed linear velocity v. This shows that although the linear velocity is relative, the angular velocity of rotation is unaffected by the movement of the observer. An observer on a rotating body cannot assume that he is at rest; as I have said before he will have feelings that belie such an assumption and, unless he is firmly secured to the body, when the speed of rotation reaches a certain value he will be flung off, whatever views he may hold on

the subject. The only relativity about this is the relative linear velocities of the different parts of the body, which produce the absolute rotation.

In the above example, if the observer were securely fastened to the wheel at the other end of the axle, both wheels would to him appear to be at rest, while the vehicle and everything else would appear to be whirling round.

The letter to which we referred at the beginning continues as follows: "Two observers A and B are placed at different distances r_1 and r_2 from a common axis, about which they have angular velocities ω_1 and ω_2. A produces an electromagnetic phenomenon due, say, to an electric charge uniformly distributed on the circle of radius r_1 and moving with A, the effects of which are observed by B. Do the results of the observation remain unchanged if the angular velocities ω_1 and ω_2 are changed without changing the difference $\omega_1 - \omega_2$? If the results vary, in what way do they change, and does the change disappear in the particular case of $r_1 = r_2$ or in the particular case of $\omega_1 = \omega_2$?"

This regards the phenomena from the point of view that seems to be very undesirable; viz., that of a rotating observer. When one places an observer on a rotor, is he supposed to ignore the centrifugal force and the giddiness that he experiences and assume that he is at rest and that the stator is rotating? Even if he were suitably secured and treated so that he would not be troubled about such details, would not his evidence be open to some doubt? Take the almost classic example in Fig. 3 of the cylindrical bar magnet rotated about its axis while an external circuit containing a galvanometer is connected by means of brushes between one end and the middle of the magnet. The galvanometer shows that a current is flowing in the circuit, and seeing that the external circuit is at rest in a magnetic field which is undergoing no change whatever either in magnitued or direction, whereas the path in the magnet is through a conductor; viz., steel, which is rotating in a magnetic field, viz., that within the magnet, which is also, speaking macroscopically, undergoing no change whatever either in magnitude or direction, no one can have any doubt as to the source of the e.m.f.

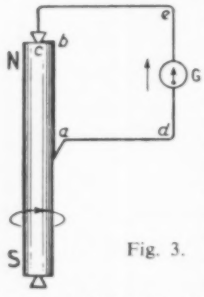

Fig. 3.

If now we imagine an observer firmly bound to the magnet, he will tell us that the magnet is at rest but that the external circuit and galvanometer are being rotated around it, thus cutting through the magnetic field and inducing an e.m.f. in the circuit. The observable phenomenon, viz., the current indicated on the galvanometer, would be the same in both cases, but the fact that he has to be fastened to the magnet shows that all his explanations are based on a false assumption, viz., that the magnet is not rotating. When the magnet is rotating all the magnetic flux emerging from the upper half of the magnet is cut by the conductor 'abc'; when the magnet is at rest and the external circuit rotating, the same flux is cut by the conductor 'adec'.

In the question put to us, a circle A is charged, say positively, while B is an observer moving about the same axis. When both are at rest B merely observes a stationary charge and its electric field—assuming that he is properly equipped for making such observations. If now A rotates, B observes the moving charge as a circular current producing a magnetic field in addition to the electric field. So long as B is at rest there is no difficulty, but if he now rotates there are real difficulties because the recognized laws of electromagnetism, like those of mechanics, are based on the assumption that the observer is not rotating. If A and B are rotating at the same speed, the charge on A appears to be at rest and the magnetic field presumably vanishes so far as B is concerned. To a stationary observer the magnetic field is still there and B is rotating in it. To go into this question more fully one would have to specify the type of body to which B was secured and the instruments with which he was equipped.

All this goes to prove that rotation is absolute; a body is either rotating or not rotating, independent of any observer. When one speaks of the speed of rotation of a body relative to that of another, one means merely $\omega_1 - \omega_2$; the values of ω_1 and ω_2 are, however, absolute and independent of any other body. From the examples that we have discussed it is obviously very undesirable to attempt the explanation of electromagnetic or mechanical phenomena from the point of view of a rotating observer. One can discuss what would happen if the rotor of a machine were at rest and the stator rotated, but this can be done without fixing a fictitious observer to the stator and running the machine in the normal way.

G. W. O. H.

MULTIPLEX TELEVISION TRANSMISSION

Sub-Carrier and Dot-Interlace Systems

By J. Haantjes and K. Teer

(Philips' Research Laboratories, N.V. Philips' Gloeilampenfabrieken, Eindhoven, Holland)

SUMMARY.—Systems for the transmission of several television signals within a single television channel are described in this article; they are based on the use of signal components which cancel out in two successive pictures.

A distinction is made between the sub-carrier and the dot-interlace systems. The typical characteristics of both systems are determined and, in particular, those characteristics which affect the separation of the signals at the receiver. Their application to colour television is considered and the conclusion is drawn that for this the sub-carrier system is to be preferred.

1. Introduction

IN the development of colour television it has been realized from the beginning that it is very desirable to be able to transmit several television signals within a single television channel. The colour and luminance of a picture element are determined by three independent values and, therefore, three sets of information have to be transmitted. From the point of view of bandwidth economy and compatibility with the black-and-white system, however, it is desirable to use the normal channel bandwidth.

Similar conditions occur in stereoscopic television and it is not impossible that in other future applications of television such a multiple transmission of television signals will be required.

This paper is concerned with this general problem of the multiple transmission of television signals. We shall discuss the simultaneous transmission of several separable television signals through a single transmission path, the bandwidth of the transmission channel being assumed to be less than the sum of the bandwidths of the individual signals.

In such a transmission system the spectra of the signals overlap and these spectra cannot, therefore, be entirely separated by means of filters. Consequently, multiple transmission involves the problem of adding to a certain television signal a second separable signal within the same band in such a way that its presence does not affect the quality of the picture corresponding to the first signal.

It is possible to solve this problem by using signal components of the transmission which are of opposite sign in two successive picture periods. Such signal components are barely visible in a television picture because, as far as the eye is concerned, the brightness modulation caused by such a signal component in one picture* cancels out when combined with the brightness modulation due to the signal component of the successive picture.

Before the use of these signal components is considered, their visual effect on a television picture will be described.

MS accepted by the Editor, April 1954

2. Signal Components Cancelling in Two Successive Pictures

One example of a signal component of the kind described is a sine wave having a frequency which is exactly an odd multiple of one-half of the picture frequency. At any instant the phase of such a sine wave is opposite to its phase one picture period earlier. A second example is a sine wave of arbitrary frequency but with a certain definite phase shift at the start of each picture period, so that the sine wave is of opposite phase in successive pictures. The cancellation effect, which results from the use of sine waves having a special frequency or phase relation to the scanning frequencies, needs detailed examination.

A sinusoidal modulation of the brightness along an individual scanning line of a picture gives the appearance of successive bright and dark areas. These areas, occurring in successive lines, frames and pictures, produce a pattern which is an unwanted one and which is, therefore, an interference pattern. To investigate the nature of this pattern we shall consider two successive pictures built up of four frames interlaced in the usual way. The relative position of bright areas, corresponding to maxima of the sine waves, will be denoted as follows:—

in the first frame by A
(odd lines, first picture)

*The usual British television nomenclature is employed in this article; i.e., the visible picture is built up from lines, frames and pictures. In American terminology, lines, fields and frames are the equivalent words. As a result, the word 'picture' is used in a double sense to mean, generally, the entire visible picture seen by the eye and, specifically, one complete cycle of the scanning process, which is usually two frames. The context prevents this double meaning from causing ambiguity.

in the second frame by B
(even lines, first picture)
in the third frame by C
(odd lines, second picture)
in the fourth frame by D
(even lines, second picture)

Now cancellation occurs if, on the odd lines, the maxima A and C alternate and, on the even lines, the maxima B and D also alternate. After two picture periods, therefore, the structure ACACAC is obtained on the odd lines and the structure BDBDBD on the even. When the resulting brightness impressions on a certain line, such as A and C or B and D, are added by the eye the effect of the sine wave is cancelled. This is illustrated by the diagram of Fig. 1.

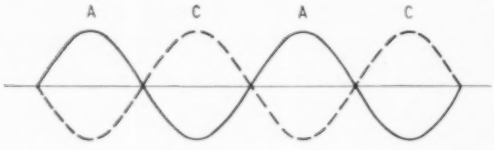

Fig. 1. *The brightness variations along a certain line in successive pictures, due to a sine wave of the requisite character.*

The addition, which occurs as a result of the integrating effect of the eye, is in reality present only when the observer concentrates upon a part of the picture without moving his eyes. Only then will successive impressions appear to be simultaneously present and only then will the effect of the sine wave be invisible.

Normally, however, the observer will not keep his attention concentrated upon a particular part of the picture, but will move his eyes in an irregular manner over the picture. Because of this movement certain stroboscopic effects can occur. If the eye is moved with the appropriate speed and in the right direction, the neutralizing condition ceases to exist and certain moving dot patterns appear in the picture. These stroboscopic effects are of the same character as those which occur in an ordinary line-interlaced television picture, with which the viewer can perceive a line pattern moving vertically.

In the case of the sinusoidal modulation of the lines the eye tends to interpret the different positions of the maxima in successive frames or pictures as a dot pattern moving in some direction. The extent to which the effect is noticeable and annoying depends mainly on the relative positions of the maxima A, B, C and D.

An example is shown in Fig. 2. The positions of the maxima of the sine wave appear visibly rather as dots and are represented in the diagram by letters. The positions of the dots on one line are displaced half the distance AA with respect to the positions on the preceding line of the same frame. The pattern shown is, of course, for the four frames of two pictures, and in the scanning process the A-dots are produced in lines 3, 5, etc., of frame 1, the B-dots in lines 2, 4, etc., of frame 2, the C-dots in lines 1, 3, 5, etc., of frame 3 and the D-dots in lines 2, 4, etc., of frame 4, after which the sequence is repeated. It can be seen that corresponding dots in adjacent lines (e.g., A and B in lines 1 and 2) occupy corresponding positions along the lines. It is plain, therefore, that the eye may interpret the four impressions A, B, C and D as one pattern A moving downwards with a velocity of one line width per frame period (vertical arrow in Fig. 2). A movement of the pattern A in a diagonal direction can also be observed (diagonal arrow). It has been found experimentally that these effects are by no means invisible.

The condition of Fig. 2 occurs when the frequency of the sine wave is an odd multiple of one-half of the line frequency. (Because of the odd number of lines in a television picture, the frequency is also an odd multiple of one-half of the frame frequency.)

The example considered is not the only possible one and there are many alternative ways of arranging the maxima on successive lines. Moreover, the disturbing effect of the sine wave is not the only effect of the moving-dot structure. 'Moiré' effects can occur in regions of the picture where the picture detail is regular and of nearly the same size as the dot structure.

Experiments have been carried out to determine the visibility and annoyance of several possible dot patterns and, without discussing them in detail, we want to draw attention to a particular pattern which shows the effects to an unusually small degree. This pattern is illustrated in Fig. 3 and the spacing between the dot positions on successive lines of one frame is still one-half of the dot spacing. The spacing between the dot positions in adjacent lines of one picture is, however, one-quarter of the dot spacing.

This pattern cannot be obtained exactly with

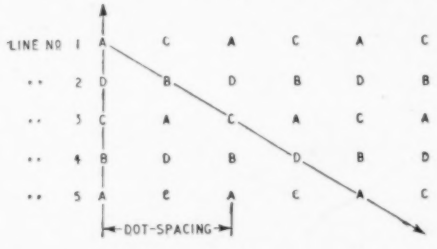

Fig. 2. *The dot pattern of a sine wave with a frequency equal to an odd multiple of half the line frequency. The arrows indicate the direction of the stroboscopic effects.*

any continuous sine wave. It can, however, be obtained by using, as before, a sine wave with a frequency equal to an odd multiple of one-half of the line frequency but this time introducing a phase shift of alternately plus and minus 90° at the beginning of each frame.

Under these conditions it can be seen from Fig. 3 that there is no longer any definite direction in which the dots A, B, C and D can appear successively. It is no longer possible to draw any straight line through successive dots, as in the case of Fig. 2. An apparent movement of the dot pattern of the type described in connection with Fig. 2 is thus impossible.

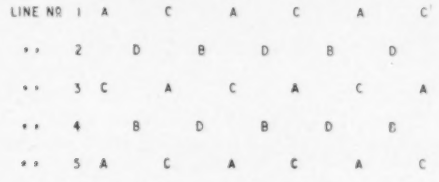

Fig. 3. *The dot pattern of an odd multiple of half the line frequency with a phase shift at the beginning of each frame of + and − 90° alternately.*

However, this is not the only condition which can produce a moving pattern; a 'second-order' moving pattern can exist. If the eye is moved in such a way that the region of attention is shifted over half the dot distance AA in a vertical or a horizontal direction during a picture period, the impression of pattern A will coincide with the impression of pattern C. Similarly, pattern B will coincide with pattern D. The observer perceives two patterns simultaneously, one caused by the coincidence of A and C and the other by the coincidence of B and D.

The frames coincide two by two, so the brightness of this second-order moving pattern will be less than that of the first-order pattern obtained with the arrangement of Fig. 2. Moreover, these second-order effects occur in both vertical and horizontal directions and are equally perceptible in all four directions. Because of this the stroboscopic effects are much less evident than with the Fig. 2 pattern.

3. The Sub-Carrier System

It is now necessary to consider how a sine wave can be used to carry a second set of information within the same band as that used for the main television signal. One possibility is to modulate the sine wave, using double-, vestigial- or single-sideband modulation, and then to employ it as a sub-carrier with the main signal.[2]

As already stated, multiples of half the picture frequency are of minor importance in the spectrum of a television signal.[1] In the case of a still picture these frequencies are not present at all, because the signal is then periodic at the picture frequency. With a moving picture, however, they do exist, but they are of relatively small amplitude; in fact, it is almost true to say that in the higher regions of the video spectrum the only frequencies present are multiples of the line frequency.

Now when a sub-carrier has a frequency which is an odd multiple of one-half of the picture frequency and is itself modulated by a television signal, the resulting sidebands have frequencies which are also odd multiples of half the picture or line frequency. This is illustrated in Fig. 4, which shows the spectrum of the modulated sub-carrier. The whole of the second television signal is, therefore, transformed into signal components which are barely visible in a reproduction of the first and main picture.

The second signal can be represented by

$$S_2 \cos \omega_p t$$

where ω_p is the angular frequency of the sub-carrier. The total signal transmitted, which is henceforth called the 'transmission signal', can be represented by the expression

$$S_1 + aS_2 \cos \omega_p t$$

where a is a constant.

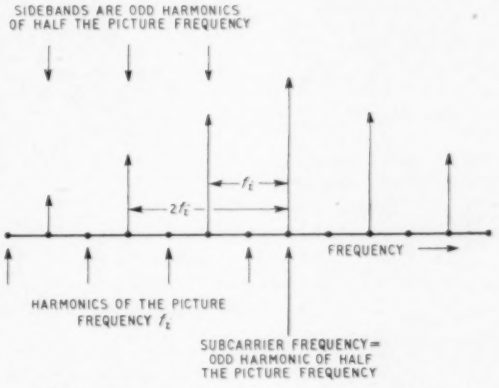
SIDEBANDS ARE ODD HARMONICS OF HALF THE PICTURE FREQUENCY
f_l
$2f_l$
FREQUENCY
HARMONICS OF THE PICTURE FREQUENCY f_l
SUBCARRIER FREQUENCY = ODD HARMONIC OF HALF THE PICTURE FREQUENCY

Fig. 4. *The spectrum of a modulated sub-carrier.*

When this signal is applied to a cathode-ray tube it will produce a visible picture corresponding to S_1 because the components of $S_2 \cos \omega_p t$ cancel each other in successive pictures. In order to obtain a picture corresponding to S_2 it is necessary to multiply the transmission signal by $\cos \omega_p t$; that is, to use synchronous detection. The result of doing this can be expressed as

$$\frac{2}{a} \cos \omega_p t \,(S_1 + aS_2 \cos \omega_p t)$$

$$= S_2 + S_2 \cos 2\omega_p t + \frac{2}{a} S_1 \cos \omega_p t$$

B

If this signal is filtered in such a way that the component $S_2 \cos 2\omega_p t$ is removed the result is

$$S_2 + \frac{2}{a} S_1 \cos \omega_p t$$

This signal has the same character as the transmission signal and can, therefore, be used to represent the signal S_2. If the constant a is chosen to be $\sqrt{2}$ the amplitude relation of the signals S_1 and S_2 is symmetrical.

If the sub-carrier is single-sideband modulated the results are similar. Apart from the advantage of smaller bandwidth of the modulated sub-carrier, single-sideband transmission has also the advantage of giving better signal-to-noise conditions at the receiver for S_2. The extreme condition for single-sideband operation is shown in Fig. 5, where the two signals have the same bandwidth f_p, which is equal to the channel bandwidth. If it is required to transmit a third signal S_3 with S_1 and S_2 a second sub-carrier modulated by S_3 could be introduced. Another method, however, is to modulate one sub-carrier by S_2 as well as S_3 in such a way that they can be separated by synchronous detection.

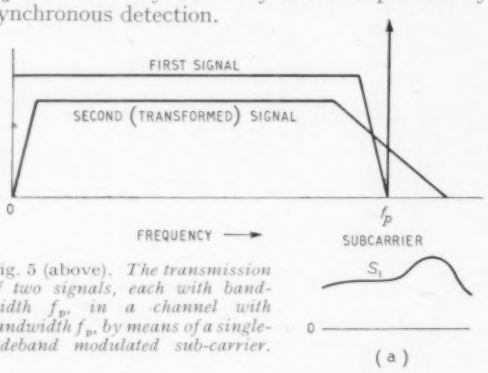

Fig. 5 (above). *The transmission of two signals, each with bandwidth f_p, in a channel with bandwidth f_p, by means of a single-sideband modulated sub-carrier.*

If S_3 is used to modulate a carrier $\sin \omega_p t$ we get

$$\sqrt{2}\, S_3 \sin \omega_p t$$

and if this is added to

$$S_1 + \sqrt{2}\, S_2 \cos \omega_p t$$

the transmission signal becomes

$$S_1 + \sqrt{2}\, S_2 \cos \omega_p t + \sqrt{2}\, S_3 \sin \omega_p t$$

When this is multiplied by $\sqrt{2} \cos \omega_p t$ we get

$$S_2 + \sqrt{2}\, S_1 \cos \omega_p t + S_2 \cos 2\omega_p t + S_3 \sin 2\omega_p t$$

and after filtering the result is

$$S_2 + \sqrt{2}\, S_1 \cos \omega_p t$$

When the transmission signal is multiplied by $\sqrt{2} \sin \omega_p t$ and the same filtering process employed the result is

$$S_3 + \sqrt{2}\, S_1 \sin \omega_p t$$

In this way the three signals can be separated and are represented by the transmission signal and the two demodulator output signals. However, single-sideband transmission of a sub-carrier modulated by two signals is not possible without introducing cross-talk[3,4] between S_2 and S_3.

This method of transmitting three signals is used in the recently-accepted colour system of the U.S.A. and usually known as the N.T.S.C. system.

4. The Dot-Interlace System

4.1. *Principles*

Instead of using a sub-carrier system of the type just described, it is possible to transmit two or more signals in one channel by employing time-division multiplex. As applied to television, this is usually known as dot interlace.

In spite of its apparent difference, one can actually detect the presence of sub-carriers having similar frequency and phase relations to the scanning frequencies, as in the case of the nominally sub-carrier system. When the method was first developed, however, it was regarded purely from the time-division point of view[4] and, initially, we also shall look at it in this way.

Suppose that a television signal S_1 has a bandwidth f_p and that this signal is multiplied by a series of pulses of infinitely-short duration. The result is a pulse series amplitude-modulated by S_1 as shown in Fig. 6.

Ideally, this signal is now fed to a perfect filter (i.e., a filter having a rectangular amplitude and a linear phase characteristic) which has a bandwidth

$$f_b = nf_p/2$$

where n is an integer.

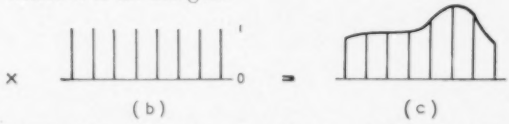

Fig. 6. *The signal S_1 shown at (a) is multiplied by a series of infinitely-short pulses (b) to produce a pulse-modulated wave (c).*

To find the output signal from this filter we consider its response to a single pulse. This response is well known[5] and is proportional to

$$\frac{\sin \omega_b t}{\omega_b t}$$

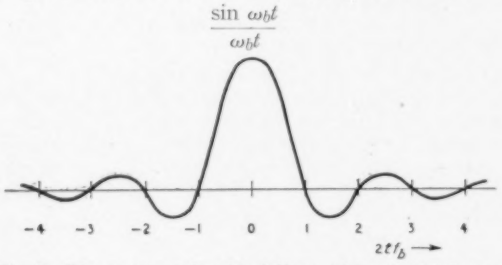

Fig. 7. *The response of an ideal filter with bandwidth f_b to a single pulse.*

which has the form shown in Fig. 7. This signal has a maximum for $t = 0$ and is zero whenever

$$t = \tfrac{1}{2}m/f_b$$

$$m = \pm 1, \pm 2, \pm 3, \text{ etc.}$$

If the signal fed to the filter is a modulated pulse train having a repetition frequency of $2f_b$, therefore, the maximum in the response to any individual pulse will coincide with the zeroes of the responses to all the other pulses. Hence, at $t = m/2f_b$, the value of the output signal depends only on the value of one single input pulse, which is the pulse having its maximum at that moment. Consequently, by multiplying the output signal from the filter by a second pulse train the original amplitude-modulated input pulse train can be reconstituted. This is shown diagrammatically in Fig. 8.

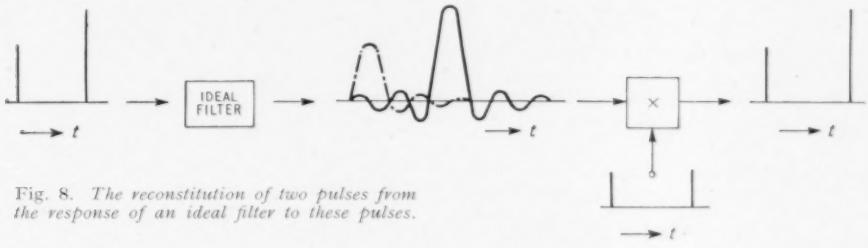

Fig. 8. *The reconstitution of two pulses from the response of an ideal filter to these pulses.*

The bandwidth of the original signal S_1 is f_p and it would appear to be necessary to have a pulse repetition frequency of $2f_p$ because a minimum of $2f_p$ independent values per second are necessary to describe[7] a signal of bandwidth f_p. However, this is not essential when S_1 is a television signal. In this case the pulse repetition frequency can be f_p only. The result is that only half as many signal values per second are transmitted as are necessary for the transmission of the entire information. In television, the transmitted time function is transformed at the receiver into a two-dimensional space function and, therefore, it is possible to add the missing values later on.

The signal values missing in a certain scan (comprising a picture period of two frames) of the picture can be transmitted during the following scan, assuming that the picture does not vary too much during one picture period. In this way the whole information content of a picture is transmitted as two series of interlaced signal values in two successive picture periods. It is plain, however, that this reduction in the number of signal values transmitted per second is only possible in television and similar techniques.

As the repetition rate is f_p, the time interval between the pulses is $1/f_p$ and the interval between two zeroes in the response of the transmission filter is

$$\frac{1}{2f_b} = \frac{1}{nf_p}, \quad \text{for} \quad f_b = \frac{nf_p}{2}$$

This interval between the pulses is, therefore, $1/n$ of the pulse interval. It follows that another $n - 1$ pulse series with suitable time shifts can be added to the filter input. The total input signal is then a pulse series with a pulse interval of $1/nf_p$. The $n - 1$ pulse series can be modulated by signals $S_2, S_3, \ldots S_n$. This is illustrated in Fig. 9 for the case of $n = 3$.

In order to separate the information, say, S_k, carried by one pulse series from the complex signal, this signal is multiplied by pulses of frequency and phase equal to those applied to S_k in the transmitter. This, of course, assumes that the delay time of the transmission path is zero; when it is not, it is only necessary to alter the phase appropriately.

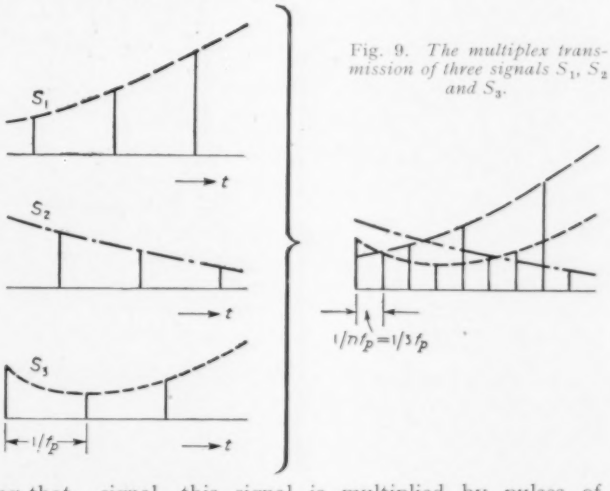

Fig. 9. *The multiplex transmission of three signals S_1, S_2 and S_3.*

The signal resulting from the multiplication is

fed to an ideal filter with bandwidth f_p, which is henceforth called the 'receiver filter'. As the bandwidth is the same as the repetition rate of the pulses, the filter response will show zeroes between

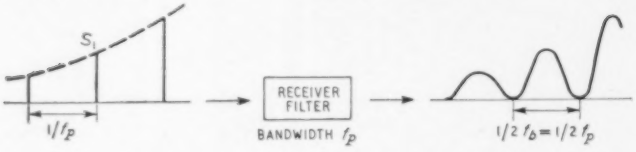

Fig. 10. *The response of the receiver filter to the reconstituted pulse series modulated by* S_1.

the maxima corresponding to two adjacent pulses. This is depicted in Fig. 10. The missing signal values occur at instants which correspond with the zeroes in one picture period, and are transmitted in the following picture period. Because of the integrating properties of the eye, the picture obtained by reproducing the output signal appears much the same as that given by the reproduction of S_k itself.

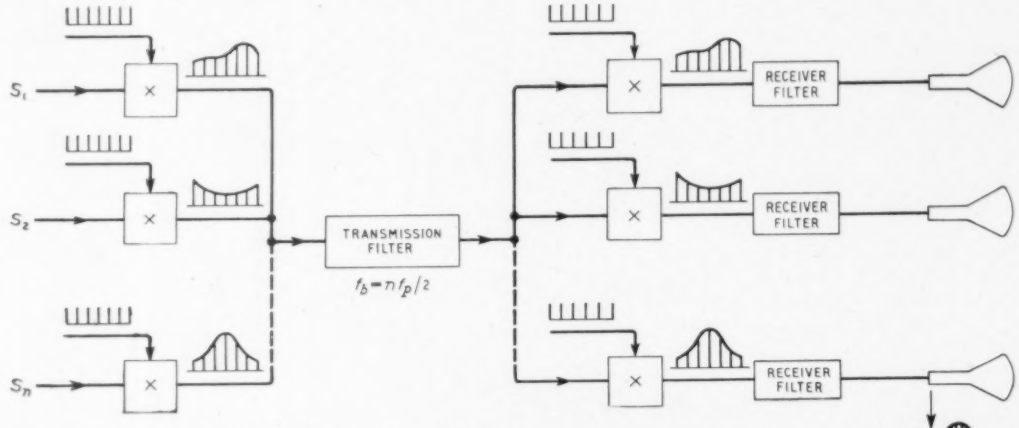

Fig. 11. *The entire dot-interlace system for n signals.*

To sum up, in a dot-interlace transmission system of the kind described, n television signals S_k ($k = 1, 2, 3, \ldots n$) with frequencies up to f_p, can be transmitted through a channel of bandwidth $nf_p/2$ and at the receiver the original picture can be reconstituted by a combination of two successive pictures. The complete system is shown schematically in Fig. 11.

4.2. *Practice*

As so far described the dot-interlace system depends upon the use of ideal filters and infinitely-short pulses, neither of which is possible in any practical system. It is, therefore, necessary to consider what happens when practicable filters and pulses are used.

The Fourier transform of a pulse of infinitely-short duration is

$$\propto \int_0^\infty \cos \omega t \, d\omega$$

and the response to it of an ideal filter with zero phase-shift is

$$\propto \int_0^{f_b} \cos \omega t \, d\omega$$

where f_b is the bandwidth of the filter.

Now suppose that, instead of the ideal rectangular characteristics, the filter has a response of the form shown in Fig. 12. This characteristic is down 50% at f_b and shows symmetry with respect to this point such that when a certain frequency $f < f_b$ is attenuated by a factor γ the frequency $f_b + (f_b - f) = 2f_b - f$ is attenuated by a factor $1 - \gamma$. The highest frequency transmitted by the filter is limited to $2f_b$.

The response of this filter to a pulse can be written as

$$\propto \int_0^{f_b} [\gamma \cos \omega t + (1 - \gamma) \cos (2\omega_b - \omega) t] \, d\omega$$

$$= \propto \int_0^{f_b} \cos \omega t \, d\omega + \propto \int_0^{f_b} 2 (1 - \gamma) \sin \omega_b t \sin (\omega_b - \omega) t \, d\omega$$

The first integral represents the response of an ideal filter. The second must be considered for those instants when $t = m/2f_b$, where m is an integer.

At these instants the second integral is zero, and so the response of the filter is then the same as that of the ideal filter. It is only at these instants that the value of the transmission signal is used to determine the information signals at the receiver and so the results with the two types of filter are the same. The filter of Fig. 12 can, therefore, be used instead of the ideal filter. The phase characteristic must still be linear, but such a filter can be realized to a good degree of approximation.[8]

It can similarly be shown that the receiver filters, of bandwidth f_p, may have amplitude characteristics like that of Fig. 12. In this case the final result is not quite the same as with ideal filters. It can be shown that in the response the output signal equals the original signal at the instants which correspond to the $2f_b$ signal values, but that slight deviations in the output signal occur between these instants.

It is now necessary to consider the effect of the pulse shape. A train of infinitely-short pulses of repetition frequency f_p can be written as a Fourier series:

$$\alpha\,[1 + 2\cos(\omega_p t + \phi) + 2\cos\{2(\omega_p t + \phi)\} + 2\cos\{3(\omega_p t + \phi)\} + \ldots.]$$

$$= \alpha\left[1 + 2\sum_{m=1}^{\infty}\cos\{m(\omega_p t + \phi)\}\right]$$

At the transmitter the signals

$$\alpha\, S_k\left[1 + 2\sum_{m=1}^{\infty}\cos\{m(\omega_p t + \phi_k)\}\right]$$

are fed to the transmission filter. The highest frequency passed by this filter is $2f_b = nf_p$ and, as S_k does not contain frequencies above f_p, the component

$$S_k\cos\{m(\omega_p t + \phi_k)\} \qquad (m > n)$$

cannot contribute to the output.

Instead of multiplication by pulses of infinitely-short duration, multiplication by the waveform

$$P_k = 1 + 2\sum_{m=1}^{n}\cos m(\omega_p t + \phi_k)$$

may be employed, therefore, with the same result.

The number of harmonics of f_p which must be present in P_k depends upon the sharpness of cut-off of the transmission filter, and decreases with an increase of sharpness. In the extreme case of an ideal filter P_k must contain harmonics up to

$$\frac{n}{2} + \frac{1}{2}, \quad n \text{ odd}$$

$$\frac{n}{2}, \quad n \text{ even}$$

In a similar manner it can be shown that in the receiver the waveform P_k can be used for separating the different signals, instead of infinitely-short pulses, to produce signals $S_k P_k$ at the input of the receiver filters. These filters, having a bandwidth f_p, cannot transmit the sidebands of the harmonics $3f_p$, $4f_p$, . . . nf_p in P_k. The output signal will, therefore, have the same form if the input signal is

$$S_k[1 + 2\cos(\omega_p t + \phi_k) + 2\cos\{2(\omega_p t + \phi_k)\}]$$

Moreover, from the component

$$2\,S_k\cos\{2(\omega_p t + \phi_k)\}$$

only the lower sideband can contribute to the filter output; and when the cut-off is sharp, this contribution is almost negligible. The output signal is then nearly S_k + the sub-carrier f_p single-sideband modulated with S_k.

If we indicate the single-sideband modulation on the sub-carrier f_p by the operator Ω we can write for the resulting signal

$$S_k + \Omega S_k$$

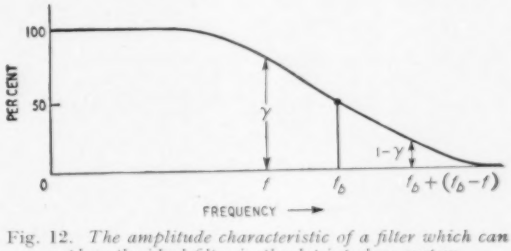

Fig. 12. *The amplitude characteristic of a filter which can replace the ideal filter in the dot-interlace system.*

4.3. *Comparison with Sub-Carrier System*

In two successive pictures the pulses must interlace; it is, therefore, necessary to reset the phase of P_k at the beginning of each picture period. This resetting becomes unnecessary, however, when there is a special relation between f_p and the picture frequency f_i, viz.,

$$f_p = (n + \tfrac{1}{2})\,f_i$$

n being an integer.

When the beginning of each picture period is indicated as $t = 0$, the condition of interlace can be formulated as requiring that the angle ϕ_k in P_k must show a difference of π radians in two successive periods. Writing

$$1 + 2\cos(\omega_p t + \phi_k) + 2\cos\{2(\omega_p t + \phi_k)\} + \ldots$$

for the first scan, we get

$$1 - 2\cos(\omega_p t + \phi_k) + 2\cos\{2(\omega_p t + \phi_k)\} - \ldots$$

in the second scan.

We shall indicate the condition of interlace by a $\pm$ sign

$$P_k = 1 \pm 2\cos(\omega_p t + \phi_k) + 2\cos\{2(\omega_p t + \phi_k)\} \pm 2\cos\{3(\omega_p t + \phi_k)\} + \ldots$$

At this point in the discussion the presence of sub-carriers in a dot-interlace system becomes clear. Instead of talking in terms of time-duration multiplexing, therefore, we can speak of using several modulated sub-carriers (ω_p, $2\omega_p$, etc.) in the transmission. Half of them (the odd harmonics of f_p) have the nature of signal components which cancel out in successive picture periods (indicated by a $\pm$ sign). When there are many signals, the description of the system in terms of sub-carriers is somewhat complicated; but where there are only a few, it may well simplify the analysis and the understanding of the transmission process.

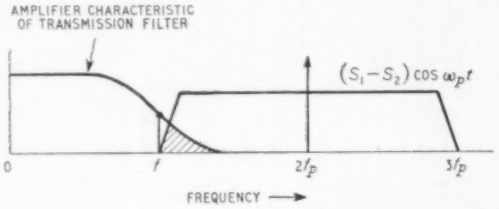

Fig. 13. *The contribution of the component* $(S_1 + S_2)$ *cos* $2\omega_p t$ *to the transmission signal of a dot-interlace system for two signals.*

The question arises, therefore, as to how the sub-carrier system described in Section 3 and the dot-interlace method described here differ from one another. In both systems signal components of the kind discussed in the foregoing are present and in both systems the observer must depend upon the integrating effect of the eye if the picture is to look right. There is, however, an important difference when the picture is considered over only one picture period. In the sub-carrier system, interference in the desired signal is related to the unwanted signals, but the whole information is contained in one picture period. In the dot-interlace system, however, the unwanted signals cause no interference with the wanted one, and defects are related to the wanted signal itself and consist of the absence of half the information in one picture period. In this system two picture periods are needed to transmit all the information.

Before we go on to consider special cases of dot-interlace, it should be mentioned that in the analysis of the dot system it has been assumed that the picture is the same in two successive scans, so that signal values which are not transmitted in the first scan can be added in the second. The analysis can, therefore, only apply exactly to the transmission of a still picture. The authors have found experimentally, however, that with a moving picture disturbing effects occur only under very special circumstances.

4.4. *Dot-Interlace with Three Signals*

In colour television three independent signals have to be transmitted, and a three-signal dot-interlace system would seem to be a suitable way of doing this. In such a system $n = 3$ and the channel bandwidth is

$$f_b = nf_p/2 = 3f_p/2$$

and so three signals, each having a bandwidth of two-thirds of the channel bandwidth, can be transmitted without cross-talk.

When used for colour television, however, such a system has the following disadvantages:—

1. All signals have a bandwidth of two-thirds of the channel bandwidth; no information at all about detail in the picture corresponding to frequencies above two-thirds of the channel bandwidth is transmitted.
2. The sub-carrier has a frequency of two-thirds of the channel bandwidth, which is rather low. Interference, in both black-and-white and colour picture reception, will be much less if the sub-carrier is chosen to be on the upper side of the channel.

Some years ago a colour television system was proposed which was very similar to the system described above. The dot-frequency was at the upper end of the channel and it was consequently free from the disadvantages referred to. It was found, however, that a serious amount of cross-talk occurred between the three signals.[9,10]

4.5. *Dot-Interlace with Two Signals*

When only two signals need be transmitted in a dot-interlace system the channel and signal bandwidths are equal,

$$f_b = nf_p/2 = f_p$$

The waveforms of P_k, where $k = 1$ and 2, are:—

$$P_1 = 1 \pm 2\cos\omega_p t + 2\cos 2\omega_p t$$
$$P_2 = 1 \mp 2\cos\omega_p t + 2\cos 2\omega_p t$$

The signal applied to the transmission filter is

$$S_1P_1 + S_2P_2 = S_1 + S_2 \pm 2(S_1 - S_2)\cos\omega_p t + 2(S_1 + S_2)\cos 2\omega_p t$$

and this comprises:—

1. The sum of the two signals.
2. The difference between the two signals modulated on to a sub-carrier cos $\omega_p t$.
3. The sum of the two signals modulated on to a sub-carrier cos $2\omega_p t$

The signals S_1 and S_2 do not contain frequencies higher than f_p, and so the component (3) can contribute to the transmission only in so far as frequencies exceeding f_p can pass through the transmission filter. An example of how this can

occur is shown in Fig. 13 and it will be seen that only a small part of the lower sideband of $2f_p$ is transmitted; this is shown hatched.

It has already been pointed out that the dot-interlace system operates correctly when the cut-off of the filter is symmetrical with respect to f_p. In such a filter the transmission of a part of the modulated sub-carrier $2f_p$ [component (3)] is always correlated with a certain attenuation of the component (1) just below f_p.

It can be shown that the error introduced by this attenuation (which manifests itself as decreased definition of the desired picture and cross-talk with the highest frequencies of the unwanted signal) is corrected by that part of component (3) transmitted by the filter and shown by hatching in Fig. 13. This agrees with the findings of Section 4.2. Component (3) contributes less and less to the transmission as the sharpness of cut-off of the filter increases, and so the error introduced by its complete omission also decreases. When component (3) is omitted the transmission signal consists of $S_1 + S_2$ (apart from the attenuation of the highest frequencies by the transmission filter) and the single-sideband modulated sub-carrier f_p with the modulation signal $S_1 - S_2$.

If the single-sideband modulation on f_p is indicated by the operator Ω the transmission signal can be written as $S_1 + S_2 \pm \Omega\,(S_1 - S_2)$. In order to obtain the signal S_1 the received signal is multiplied by

$$P_1 = 1 \pm 2\cos\omega_p t + 2\cos 2\omega_p t$$

and the result is passed through a filter with a bandwidth f_p. When the cut-off of the receiver filter is not too flat the term $2\cos 2\omega_p t$ is of only minor importance, because both the received signal and the receiver filter have a bandwidth f_p and the situation is the same as in Fig. 13. If this term is omitted, therefore, the input to the receiver filter will be

$$\{S_1 + S_2 \pm \Omega\,(S_1 - S_2)\}\ (1 \pm 2\cos\omega_p t)$$

Keeping in mind that ΩS indicated the sub-carrier f_p single-sideband modulated with S, we can write for the filter output

$$\begin{aligned} &S_1 + S_2 \pm \Omega(S_1 - S_2) \pm \Omega(S_1 + S_2) + S_1 - S_2 \\ &\quad = 2(S_1 \pm \Omega S_1) \end{aligned}$$

In this expression the attenuation of the highest frequencies of S_1 and S_2 has again been neglected.

This analysis in terms of modulated sub-carriers shows the same property as the dot-interlace system, treated in Section 4.1 in terms of pulse responses. The original signals can be obtained at the receiver, but with the addition of an undesired signal which is, however, related only to the desired signal itself.

It is now possible to answer the question, what is the difference between a modulated sub-carrier system and a dot-interlace system. In the first case the transmitted signal is

$$S_1 \pm \Omega S_2 \quad \text{(See Fig. 5)}$$

and in the second case the signal is

$$S_1 + S_2 \pm \Omega(S_1 - S_2)$$

It can be seen, therefore, that, neglecting certain effects, a dot-interlace system is analogous to a modulated sub-carrier system. The dot-interlace system can be considered as a sub-carrier system in which the normal transmitted signal consists of the sum of the two original signals and in which the difference between the two signals is modulated on the sub-carrier.

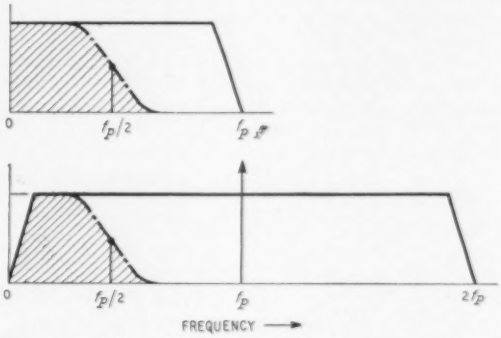

Fig. 14. *The spectrum of the signal S when transmitted by a dot-interlace system for a single signal. -·-·-· = amplitude characteristic of the transmission filter.*

4.6. *One-Signal Application of Dot-Interlace*

For the sake of completeness some mention should be made of the transmission of one signal only by dot-interlace. The channel bandwidth need be only $f_p/2$ for a signal containing frequencies up to $f_p.(n = 1)$. The pulse waveform is

$$P = 1 \pm 2\cos\omega_p t$$

and the transmitted signal is

$$S\,(1 \pm 2\cos\omega_p t)$$

as far as it passes through the transmission filter of bandwidth $\frac{1}{2}f_p$. It thus contains one-half of the spectrum of S and one-half of the lower sideband of the modulated sub-carrier f_p.

The lower half of the spectrum of S is, therefore, transmitted in the normal way and the upper half as a part of the lower sideband of f_p, as shown in Fig. 14.

At the receiver the transmission signal is multiplied by P and the signal produced at the output of the receiver filter of bandwidth f_p is

$$S \pm \Omega S$$

(To be concluded.)

DIRECT-COUPLED TRANSITRON

By F. A. Milne, A.M.I.E.E., and E. J. Miller, B.Sc.(Eng.), A.M.I.E.E.

(*P.O. Research Station*)

PENTODE-VALVE circuits using a capacitor coupling between the screen grid and suppressor grid have now become familiar in many applications and are usually known as transitrons. This type of coupling, using the mutual effect of both grids upon the anode-current characteristic, is useful if repetitive operations are required. By using direct coupling (i.e., with a resistance network) the valve may be employed for semi-static operation and in this form it has many practical applications.

The features of this type of coupled circuit are less well known and only one reference to it has been traced.[1] However, this reference does not elaborate on its performance and gives no indication of its possible usage.

Some possible applications are:—

(1) As a detector of direct voltage change; e.g., signal-level alarms, amplifier-gain alarm and light-sensitive cells.

(2) As a detector of alternating-voltage change; e.g., peak suppressors and overload alarms.

(3) As a rectifier of alternating voltages which exceed a given level; e.g., limited amplifiers.

Basic Circuit for D.C. Operation

The essential circuit is shown in Fig. 1. It will be seen that the screen-grid supply is obtained from the h.t. line, and that the suppressor-grid voltage is obtained by dividing the voltage between the screen grid and a negative supply in some fixed ratio. Usually the resistors used for obtaining the suppressor-grid voltage are such that their current does not materially alter the screen-grid voltage.

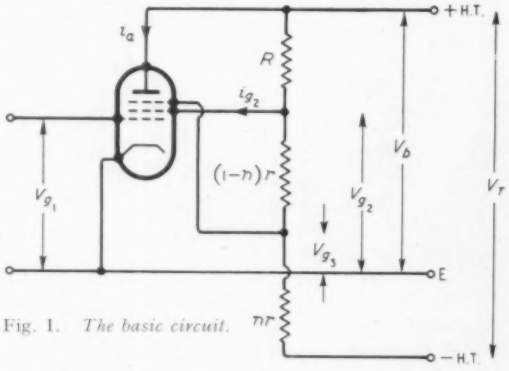

Fig. 1. *The basic circuit.*

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This is not a necessity, but it does simplify calculations.

Fig. 2 shows the variation of anode current obtained by altering the control-grid voltage. The

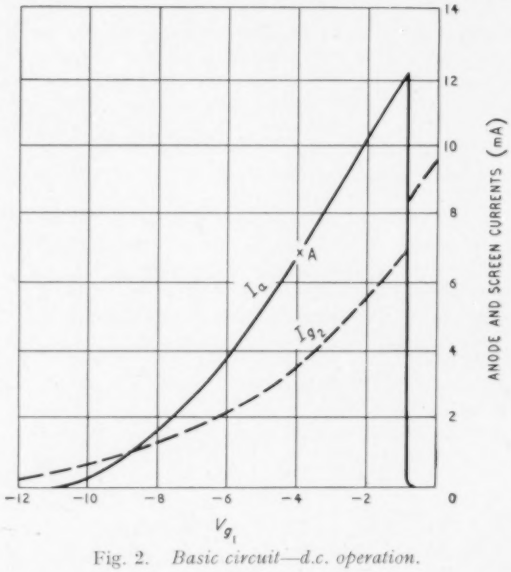

Fig. 2. *Basic circuit—d.c. operation.*

characteristic is similar to the normal pentode anode-current versus grid-voltage curve, but has a sharp cut-off at a control-grid voltage of approximately $-0{\cdot}8$. This sharp cut-off is the most useful feature of the circuit and is used in one form or another in all the applications to be discussed.

The screen-grid current characteristic is also shown in Fig. 2 and it will be seen that at the anode-current cut-off the screen-grid current is not only maintained but has an abrupt rise at the actual cut-off point.

It is interesting to note that the circuit has three stable states: (i) when the valve is completely cut off, (ii) when the valve is working normally such as at point A in Fig. 2, and (iii) when the anode current only is cut off.

The rapid anode-current cut-off is achieved by transitron action in the following manner. Consider a reduction in the negative control-grid voltage. There will then be an increase in the cathode current; in the normally-connected pentode this would result in an increase in anode

current and a proportionately smaller increase in screen-grid current. However, in this case the increase in screen-grid current reduces the screen-grid voltage due to the series resistor R. Now the suppressor-grid voltage, being derived from the screen-grid voltage by a fixed potential divider, also falls and diverts more of the anode current to the screen grid. Over the range of high negative control-grid voltage to the anode cut-off point, the suppressor grid is maintained positive and therefore has little influence. It is only when the suppressor grid becomes negative that it becomes effective. When it does become effective the diversion of the anode current to the screen grid is cumulative and a rapid anode-current cut-off ensues. Clearly, this cut-off is better defined if the valve used has a suppressor grid with a high degree of control over the anode current. This implies that the valve should have a short suppressor-grid base. In the curves shown, a CV 329 valve was used throughout.

Although a complete mathematical analysis of the operation is complex, a graphical method enables the essential features of operation to be deduced. When designing such a circuit it is usual to require the anode-current cut-off to occur before the control grid begins to draw current, and also that the anode current before cut-off should be at a maximum.

Let V_b = h.t. line voltage
V_t = total voltage between the h.t. line and the negative supply
V_{g_1} = control-grid voltage
V_{g_2} = screen-grid voltage
V_{g_3} = suppressor-grid voltage
i_a = anode current
i_{g_2} = screen-grid current
i_k = cathode current
R_k = cathode resistance

nr and $(n-1)r$ are the values of the resistors supplying the suppressor-grid voltage from the screen-grid supply where n is less than unity.

From Fig. 1 it will be seen that:

$$V_{g_3} = n(V_t - i_{g_2}R) - (V_t - V_b)$$
$$= V_t(n-1) + V_b - ni_{g_2}R$$

where the current drawn by nr and $(1-n)r$ is neglected. If it is assumed that for anode-current cut-off V_{g_3} must be zero, then in the above equation, values of n, i_{g_2} and R must be determined. This can be done by drawing the valve characteristics of screen-grid current against control-grid voltage for various values of screen-grid voltage, as shown in Fig. 3. Load lines for various values of R are also shown and they are drawn for an h.t. supply of 250 V. The minimum permissible value of R can be determined such that the valve does not exceed its rated screen-grid dissipation. The other limiting factor is the control-grid current which will occur at a control-grid potential of approximately $-0{\cdot}7$ V. If loading of the control-grid circuit is to be avoided, the anode-current cut-off must be at approximately -1 V on the control grid and hence the value of i_{g_2} at cut-off is determined. Substituting these values of i_{g_2} and R in the above equation with $V_{g_3} = 0$ a suitable value of n can be assigned.

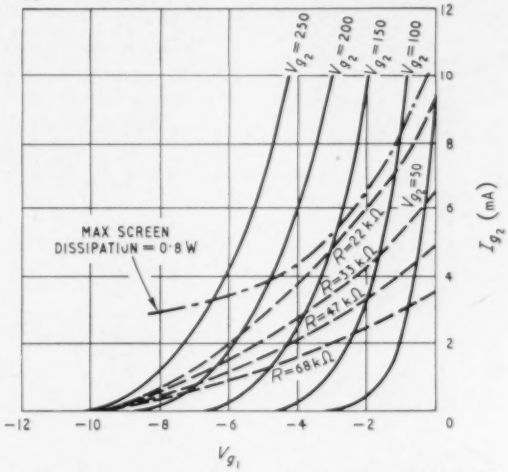

Fig. 3. *Screen-grid current characteristics of valve Type CV 329; $V_a = 200$; $V_{g3} = 0$.*

Basic Circuit with Feedback

The circuit so far discussed has an anode-current versus control-grid voltage characteristic of the form shown in Fig. 2, and this is true for any method of setting the control-grid voltage. If, however, a resistance is included in the cathode circuit this does not apply.

Consider a circuit such as that shown in Fig. 4. If the control-grid voltage with respect to earth is decreased from a high negative value, the anode-current characteristic will be similar to that of the original circuit, but the control-grid voltage will have to be increased to overcome the voltage appearing across the cathode resistance. This feedback voltage will be $i_kR_k = (i_a + i_{g_2})R_k$. This voltage will be zero when i_k is zero and large when i_k is large, thus the effective grid base of the valve can be controlled by the feedback.

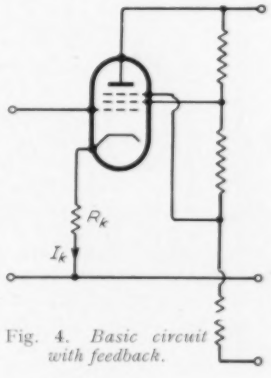

Fig. 4. *Basic circuit with feedback.*

Another interesting and important feature is the fact that after anode current has been cut off by the application of a sufficiently high value of positive control-grid bias, reducing the bias will not restore current at the same point at which it cut off because i_a is now zero and the feedback voltage is only that derived from i_{g_2}. This feature gives the circuit a hysteresis factor which can be most useful.

Fig. 5 shows the characteristic of a typical circuit with the point at which cut-off occurs and the point where it restores. If the valve is so biased that it lies within these two points it is then in a state of unstable equilibrium and may be operated as a 'trigger'. This can be called the bias for self-locking operation shown as point B in Fig. 5.

Of course, if desired, the valve may be biased so that it is not self-locking and then the hysteresis factor may be employed to control the operate and restore characteristics of the circuit with respect to time or some other function.

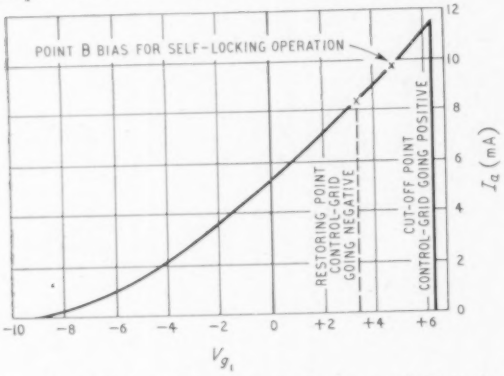

Fig. 5. *Basic circuit with feedback; anode-current—grid-voltage characteristic.*

Detector of Direct Voltage Change

(i) *Amplifier Speech-Level Alarm*

The circuit of Fig. 4 has been used as a speech-level alarm for an amplifier. Fig. 6 shows the actual circuit used. It will be noted that relays have been included in the anode and screen-grid circuits of the valve.

The requirement was to provide an alarm on a distribution amplifier for the Post Office Speaking Clock service. It was specified that an alarm should be given if the amplifier output level was reduced by 5 db but not if it was reduced by 4 db. Similarly, an alarm should be given if the output level was increased by 10 db. The overload and underload alarms were to be separately indicated. Clearly this underload-alarm condition was difficult to define as the Speaking Clock announcement is of an indeterminate character and has many quiet periods.

The output of the amplifier was rectified by a conventional full-wave rectifier and smoothed, then applied to the control-grid of the transitron shown in Fig. 6. It will be seen from the circuit characteristics that the transitron sharp cut-off enables the strict underload-alarm condition to be met, with the additional advantage that the alarm is not self-restoring. Thus, where the amplifier is in a border-line case of underload, the alarm does not follow the speech level and operates only when the speech is at a minimum. In other words, the underload alarm is decisive and positive in action.

In normal conditions relays A and B are so arranged that contacts A_1 and B_1 are closed with B_2 open; in the case of an underload alarm, relay A releases and the alarm is given by contact A_1 since contact B_2 is open. In the case of an overload, the alarm is given through B_1, while the underload alarm is not given since B_2 has closed.

A manual reset key is provided which releases the suppressor grid from its negative potential and allows the valve to conduct again.

The circuit is, of course, self-alarming since under normal operating conditions both relays are held.

(ii) *Amplifier Gain Alarm*

In another case, the circuit of Fig. 4 has been used for an amplifier gain or distortion alarm. Here the output is continuously compared with the input of the amplifier, the circuit of Fig. 7 being used to perform this operation. It will be seen that the amplifier termination has a small resistor included in series. The voltage across this resistor is opposed by a similar voltage derived

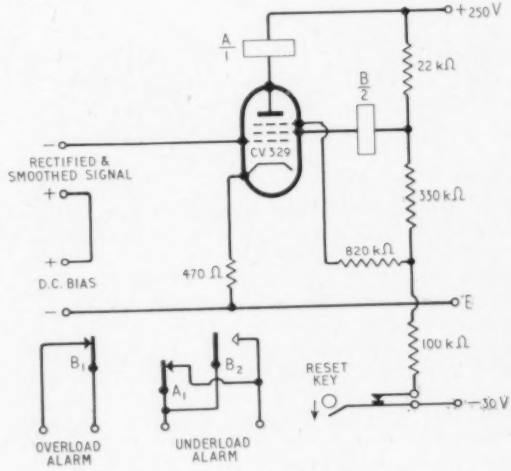

Fig. 6. *Amplifier speech-level alarm.*

from the input of the amplifier and applied across the primary of a transformer. There is a buffer cathode-follower in the case of the voltage obtained from the input of the amplifier.

The difference between these two voltages appears across the secondary of the transformer and is rectified and doubled. Before this direct voltage is applied to the transitron it has some degree of smoothing applied.

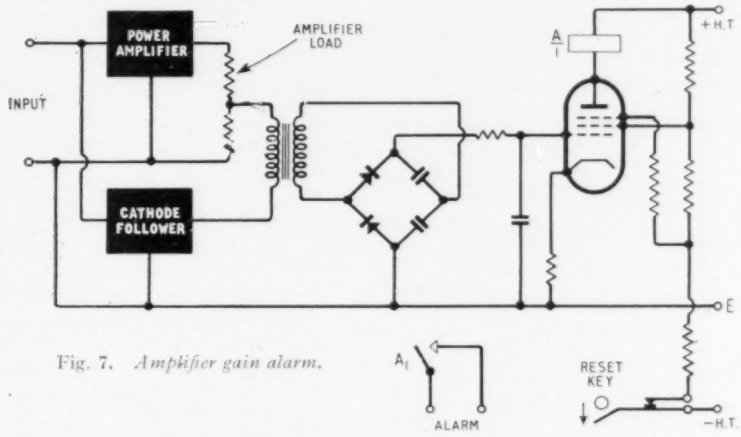

Fig. 7. *Amplifier gain alarm.*

As before, the circuit under fault conditions locks out and has to be reset manually. In this case, however, the alarm is given if the amplifier gain varies by more than a given amount about a mean value.

This type of alarm has the advantage that it can be made fast in action yet independent of the signal-input level. It is important to note, however, that the gain variation which can be tolerated is dependent on the signal level. Any undue distortion or hum arising within the amplifier will be immediately detected.

(iii) *Light-Sensitive Cells*

The circuit of Fig. 1 could be used following a photo-electric cell where an on-off operation is required. Normally where a valve follows a photo-electric cell, sufficient voltage must be derived from the photocell load to drive the valve negatively into cut-off in order to release an anode relay. Using the transitron and a photocell with a cathode load, the transitron could be driven positively into cut-off which would require less driving voltage since the transitron has a better-defined positive cut-off than the negative cut-off of a normal valve circuit. The attendant advantages would be a reduced photocell current in the working condition and less light required on the photocell cathode for efficient operation.

Basic Circuit for A.C. Operation

Consider the valve of Fig. 1 biased to the point A in Fig. 2. Now if an a.c. signal is also applied to the control grid the valve will function as a normal pentode amplifier with a resistive load provided that the control-grid excursions are not large enough to run into the point of anode-current cut-off. If the positive excursions are made so large that anode-current cut-off is encountered, it will be evident that considerable distortion of the output waveform results. The important feature of this is that a large distortion takes place at a well-defined point.

Fig. 8 shows the type of distortion that takes place and, in the case of a sine wave, it is almost a frequency doubling that occurs.

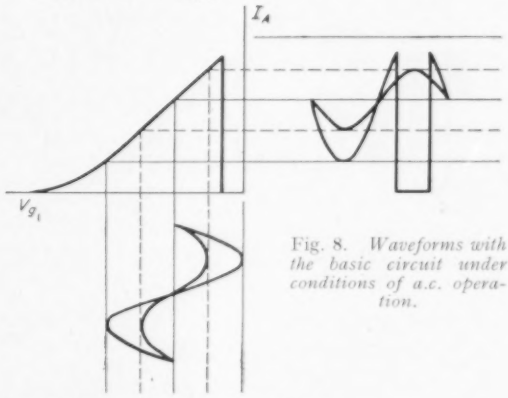

Fig. 8. *Waveforms with the basic circuit under conditions of a.c. operation.*

Feedback in Circuit for A.C. Operation

Consider, also, the effect of including a cathode resistor in the circuit. If an a.c. signal is applied, the circuit will behave as before for low-amplitude a.c. signals but where anode-current cut-off occurs, the valve, by suitable biasing (e.g., point B in Fig. 5) can be made to switch off and remain off until manually restored.

Alternatively, the bias point can be so arranged that on increasing the a.c. signal applied, the valve will switch off when anode cut-off is encountered and remain off until the a.c. signal is increased to the point where the anode current restores.

Detector of Alternating Voltage Change

Considering the a.c. operation of the circuit and the ability of the valve to cease operation when a high peak value of signal is encountered, it would appear eminently suitable as a peak-limiting stage. Some advantages would be obtained over the conventional diode limiter, namely, the level at which limiting takes place would be better defined and limiting could be performed at lower levels than would be possible with diodes.

Another possible use is an a.c. signal alarm where an alarm is required if an a.c. signal exceeds a given value. Such an alarm might be required for a thermostatically-controlled oven where the control is exercised by an a.c. bridge. The transitron could be biased for self-locking operation and give an alarm when the bridge detector current exceeded a normal value.

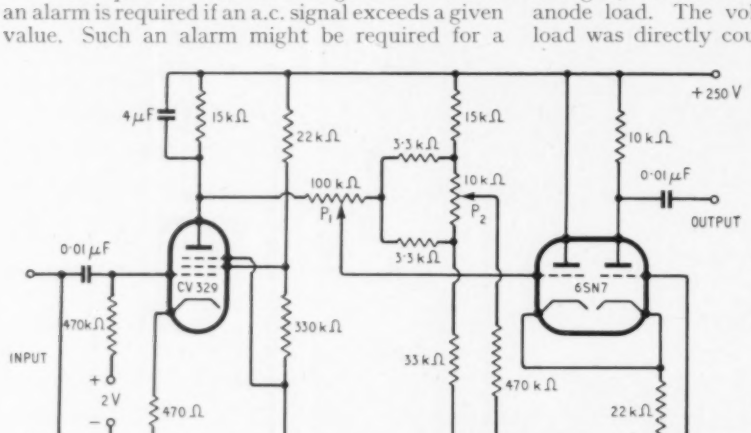

Fig. 9. *Experimental limited amplifier.*

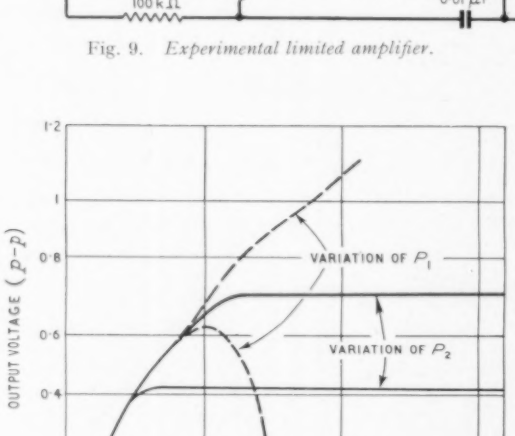

Fig. 10. *Characteristics of experimental limited amplifier.*

Rectifier of Alternating Voltages

In an attempt to utilize one of these interesting characteristics of the transitron, an experimental limited amplifier was built. In this case, as shown in Fig. 9, the transitron had a capacitor across its anode load. The voltage appearing across this load was directly coupled to the control grid of one half of a 6SN7. The other half of the 6SN7 was used as a triode amplifier and the two halves were coupled by a common cathode load.

Now, clearly, for a small value of a.c. signal the transitron behaves as a normal pentode and the charge on the anode capacitor remains constant as the a.c. signal is increased. Hence the working point of the triode amplifier remains constant and the input-output characteristic of this stage remains normal. However, when cut-off of the transitron occurs, the charge on the anode capacitor is suddenly reduced and the working point of the triode amplifier is rapidly changed.

Two potentiometers have been included in the coupling of the 6SN7 to the transitron. Potentiometer P_1 determines the amount of the voltage

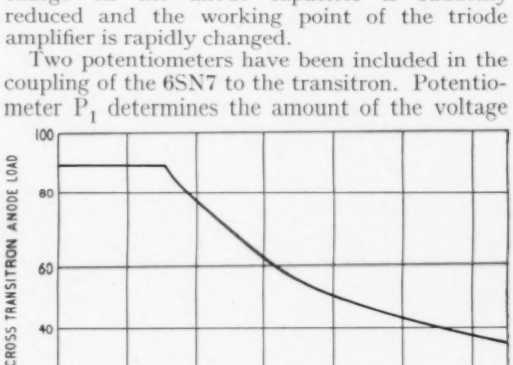

Fig. 11. *Experimental limited amplifier; control-voltage characteristic.*

change across the transitron anode load which actually reaches the grid of the 6SN7. Potentiometer P_2 determines the working point of the triode amplifier. Thus, in effect, P_1 decides the amount of level control applied and P_2 the level at which the control is applied.

Fig. 10 shows the characteristics that have been obtained with this amplifier, and it will be seen that by suitable adjustment of the potentiometers a limiting characteristic may be obtained which functions over a wide range of input levels. Of course, the experimental amplifier shown has no gain, but a considerable loss, due to the need of obtaining the correct levels for the limiting stage and for the transitron. Fig. 11 shows the change of voltage experienced at the transitron anode and clearly illustrates the well-defined and rapid change that occurs when the anode-current cut-off is reached. By a suitable matching of this characteristic with that of a controlled valve it should be possible to obtain an almost ideal limiting characteristic. In a practical amplifier the transitron would derive its signal from the latter stages while controlling an earlier stage.

Other similar applications may come to mind such as an efficient frequency doubler or as the control element in a stabilized power supply with a flexible regulation characteristic.

Conclusions

Some possibilities of a little-known circuit have been explored but many more applications remain. In practice, the circuit requires careful design to overcome variations encountered between individual valves, but an exceedingly sharp trigger action can be produced.

The sensitivity of the circuit gives many circuit economies over conventional arrangements.

REFERENCE

[1] H. J. Reich, "Theory and Applications of Electron Tubes", H. J. Reich, Chapter 10, p. 351.

IMPEDANCE TRANSFORMATION BY FOUR-TERMINAL NETWORKS

Use of Graphical Methods

By S. Fedida, B.Sc., A.C.G.I., A.M.I.E.E.

(Concluded from p. 214, August issue)

4. Graphical Determination of Network Parameters

The investigation of Sections 2 and 3 has given us a good insight into the way in which the input impedance of a four-terminal network varies when the load impedance is made to change in certain specified ways. For example, if the phase angle of the load impedance remains constant, then the locus of the input impedance, either in the input-impedance or in the input reflection-coefficient diagrams is a circle. This is the case when, in high-frequency transmission lines, the load impedance is made up of a movable short circuit.

We shall now make use of this knowledge in the determination of some of the more important network parameters. These parameters, with certain exceptions, will not and, in general, cannot, be measured directly. Instead we shall make measurements of input impedance for certain known values of load impedance, or for unknown load impedances possessing certain known properties, e.g., pure reactances.

A purely reactive load impedance in high-frequency systems is a short-circuiting plunger, the position of which may be varied. A constant-conductance load can be obtained by means of a length of transmission line, terminated resistively at the far end, with a short-circuiting stub in shunt somewhere in front of it. A variable load of constant magnitude of reflection coefficient can be made up of a mismatched termination placed at the end of a line stretcher or, alternatively, it can be the input impedance of a matched attenuator, the output terminals of which are connected to a movable short circuit.

It is sufficient for the determination of any four-terminal network constants to measure, say, three out of the four open-circuit and short-circuit input impedances. Often, however, it is quicker to determine the required parameters directly.

The determination of open-circuit input impedances, in high-frequency systems, is none too easy since it involves the realization of a

good open circuit. This is usually obtained with the aid of a $\lambda/4$ length of transmission line shorted at the far end. For accurate measurements, short-circuiting pistons must be provided with micrometer screws and, in the case of hollow waveguides, a preliminary measurement of the correct transmission-line wavelength must be made.

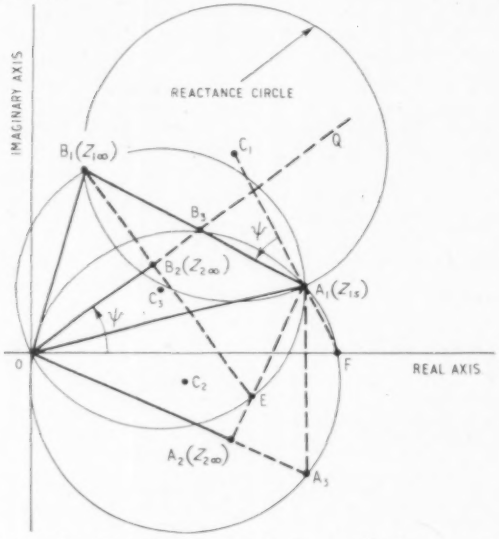

Fig. 26. *Input-impedance diagram. Construction of open-circuit input impedances ($Z_{1\infty}$, $Z_{2\infty}$), given short-circuit input impedances (Z_{1s}, Z_{2s}) and reactance circle from terminals* 1.

In view of these difficulties, it is preferable, especially when the necessary equipment is not available, to determine, instead, the short-circuit input impedances and the reactance circles. From this data the open-circuit input impedances can be deduced. This method has also the advantage of being self-checking and of smoothing out measurement errors, since the input-impedance points are known to lie on a circle.

4.1. *Determination of Open-Circuit Input Impedance*

4.1.1. *Given short-circuit input impedance and reactance circle from terminals* 1

This construction is of interest if both pairs of network terminals are accessible for measurements.

Let C_1 (Fig. 26) be the centre of the reactance circle, taken from terminals 1, A_1, B_1, A_2 and B_2 be points representing Z_{1s}, $Z_{1\infty}$, Z_{2s}, and $Z_{2\infty}$, respectively. Given C_1, A_1 and A_2 it is required to determine points B_1 and B_2.

If F is the intersection of radius C_1A_1 with the real axis and B the intersection of OB_2 with A_1B_1, the angles marked as ψ in the figure are equal and consequently points O, B_1, A_1 and F are on a common circle. This property follows from the fact that to the imaginary axis and to line OQ in the load-impedance diagram, correspond the reactance circle and line A_1B_1 in the input-impedance diagram. Hence the angle $\left(\frac{\pi}{2}-\psi\right)$ at which these two pairs of lines intersect is the same, since it is preserved in the transformation (see Sect. 2.1).

Let A_3 be the intersection of OA_2 with circle C_2. The two triangles OB_1B_3 and OA_1A_3 are similar since

$$\frac{Z_{1s}}{Z_{2s}}=\frac{Z_{1\infty}}{Z_{2\infty}} \qquad \ldots\ \ldots\ \ldots\ \ldots \qquad (40)$$

and their angles at B_3 and A_3 subtend equal arcs on circle C_2. Consequently their angles at B_1 and A_1 are equal and the circle of centre C_3, passing through O, B_1 and A_1 is tangent at A_1 to A_1A_3.

Finally, if E is the intersection of A_1A_2 and B_1B_2, it must be on circle C_3 since, according to equation (40) the angles OB_1B_2 and OA_1A_2 are equal.

Using the above properties, we can, given C_1, A_1 and A_2, determine B_1 and B_2 as follows. First construct circle C_2 passing through O, A_1 and F, the latter point being the intersection of C_1A_1 with the real axis. Produce OA_2 to A_3, and draw circle of centre C_3, through O and A_1 and tangent at A_1 to A_3A_1. The intersection of C_3 and C_1 provides B_1, the point representing $Z_{1\infty}$.

To determine B_2, find B_3 at the intersection of A_1B_1 with circle C_2, and E, at the intersection of

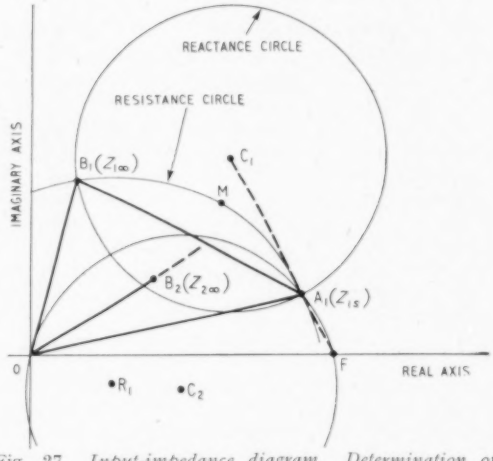

Fig. 27. *Input-impedance diagram. Determination of open-circuit input impedances ($Z_{1\infty}$, $Z_{2\infty}$), given short-circuit input impedance (Z_s) input impedance due to a load impedance of unity (point M) and reactance circle, all measurements being taken from terminals* 1.

A_1A_2 with circle C_3. B_2 is the intersection of B_1E with OB_3.

4.1.2. *Given short-circuit input impedance, input impedance due to a known resistive load and reactance circle, all measurements being taken from terminals* 1

It is not always convenient, or possible, in practice, to take measurements from both ends of the network, and in such cases the constructions given above cannot be used. An alternative method, making use of measurements taken from one end only of the network is given here. Let C_1, A_1 and M_1 (Fig. 27), be the points representing the centre of the reactance circle, the short-circuit input impedance and the input impedance due to, say, a load impedance of unity, assuming all measurements to be made from terminals 1.

Point B_1, representing the open-circuit input impedance from terminals 1, is found at the intersection of the reactance circle and the resistance circle (which represents the real axis). The latter is known to pass through A_1 and M and to be orthogonal to the reactance circle. Its centre R_1 is thus simply determined.

If F is the intersection of C_1A_1 with the real axis, it has been proved, in 4.1.1 above, that the circle through A_1, F and O intersects A_1B_1 at B_3, so that the vector $Z_{2\infty}$ lies along OB_3. The magnitude of $Z_{2\infty}$ is, according to equation (9), given by the ratio MB_1 to MA_1, hence B_2 is easily found. Z_{2s} can now be determined by constructing a triangle OA_1A_2 (Fig. 26) similar to triangle OB_1B_2.

4.1.3. *Case of symmetrical networks. Given short-circuit input impedance and reactance circle*

When the given network is known to be symmetrical, the construction given in 4.1.1 simplifies considerably. Referring to Figs. 26 and 28, we see that in this event point B_2 coincides with B_1 and A_2 with A_1. It follows that the circle of centre C_2, passing through O, A_1 and F, intersects circle C_1 at B_1, the point representing the open-circuit input impedance.

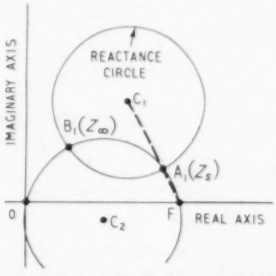

Fig. 28. *Input-impedance diagram. Construction of open-circuit input impedance (Z_∞), given short-circuit input impedance (Z_s) and reactance circle, for symmetrical networks.*

4.2. *Determination of Input Open-Circuit Reflection Coefficients*

When the measured data is given in the Smith Chart it is possible to determine the open-circuit input reflection coefficients using the methods indicated below.

4.2.1. *Given short-circuit input reflection coefficients and reactance circle from terminals* 1

The construction in the K diagram follows closely that in the impedance diagram explained in 4.1.1.

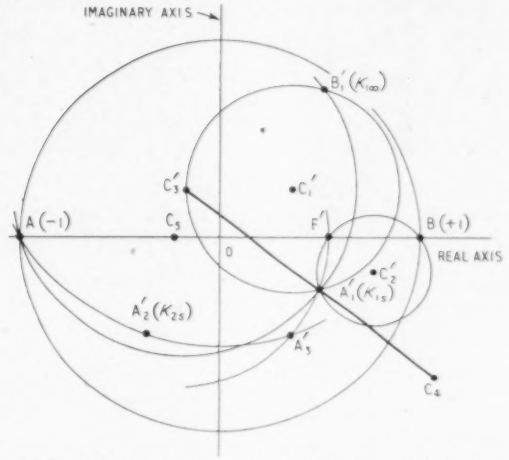

Fig. 29. *Input reflection-coefficient diagram. Determination of open-circuit reflection coefficients ($K_{1\infty}$, $K_{2\infty}$), given short-circuit reflection coefficients (K_{1s}, K_{2s}) and reactance circle from terminals* 1 *(circle of centre C'_1).*

Let C'_1, A'_1 and A'_2 (Fig. 29) represent the centre of the reactance circle, taken from terminals 1, and the short-circuit input reflection coefficients from terminals 1 and 2 respectively. The circle corresponding to the radius C_1A_1, in Fig. 26, passes through A'_1, which corresponds to A_1 and B, which corresponds to the points on C_1A_1 situated at infinity, and it has the property of being orthogonal to the reactance circle. Its centre is easily found at C'_2 and its intersection with the real axis is F', which corresponds to F.

Similarly A'_3, the point corresponding to A_3, is found at the intersection of a circle of centre C_5, corresponding to circle C_2, passing through F', A'_1 and A, and a circle, corresponding to the line OA_2, passing through A, B and A'_2. The straight line A_1A_3 becomes the circle of centre C_4, passing through A'_1, A'_3 and B. Finally, to circle C_3 corresponds a circle of centre C'_3, passing through A and A'_1 and touching circle C_4 at A_1. B'_1, the open-circuit input reflection coefficient, is found at the intersection of circles C'_1 and C'_3.

The same general method can be used for the determination of B'_2, the point corresponding to B_2.

4.2.2. *Given short-circuit input reflection coefficient, input reflection coefficient due to a known resistive load and reactance circle, it being assumed that all measurements are taken from terminals* 1

In this case we follow closely the method indicated in **4.1.2**.

We first draw the resistance circle of centre R'_1 (Fig. 30), which passes through A'_1, M'_1 and is orthogonal to the reactance circle. The second

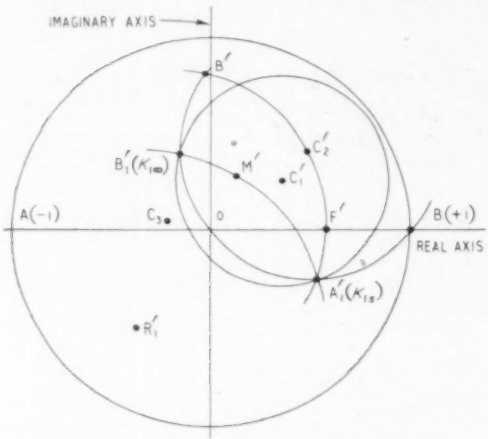

Fig. 30. *Input reflection-coefficient diagram. Determination of open-circuit input reflection coefficient* ($K_{1\infty}$), *given short-circuit input reflection coefficient* (K_{1s}), *input reflection coefficient (point M'), due to unity load impedance and reactance circle (circle of centre* C'_1), *all measurements being taken from terminals* 1.

intersection B'_1, with the reactance circle, provides the open-circuit reflection coefficient from terminals 1.

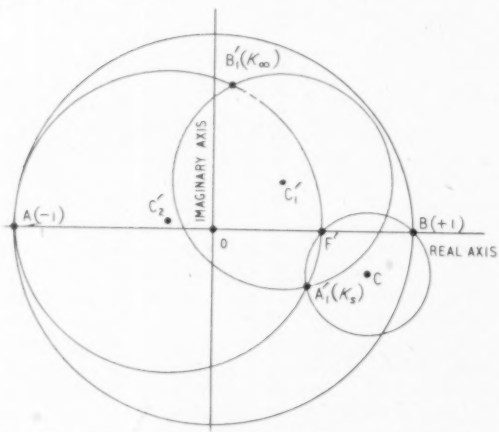

Fig. 31. *Input reflection-coefficient diagram. Determination of open-circuit input reflection coefficient* (K_∞), *given short-circuit input reflection coefficient* (K_s) *and reactance circle (circle of centre* C'_1), *for a symmetrical network.*

B'_3, the point corresponding to B_3, in Fig. 27, is found at the intersection of circle C_3, through A, A'_1 and F'_1, the latter point being determined as in the above paragraph, and circle C'_2, corresponding to line A_1B_1, which passes through A'_1, B'_1 and B. B'_2, the point representing $K_{2\infty}$, is on the circle through B'_1, A and B, which corresponds to the line OB_3. The magnitude of $Z_{2\infty}$ can now be readily determined from equation (**20**), all the terms in that equation being easily measureable on the chart.

4.2.3. *Case of symmetrical networks. Given short-circuit reflection coefficient and reactance circle*

Following the method of **4.1.3**, we first determine the circle of centre C (Fig. **31**), which corresponds to the radius C_1A_1 (Fig. **28**). Circle C is orthogonal to circle C'_1, the reactance circle, and it passes through A'_1 and B. Its intersection F', with the real axis, corresponds to point F. To circle C_2, in Fig. 28, corresponds circle C'_2, passing through F', A'_1 and A. The intersection of C'_2 with C'_1 provides B'_1, the point representing K_∞, the open-circuit input reflection coefficient.

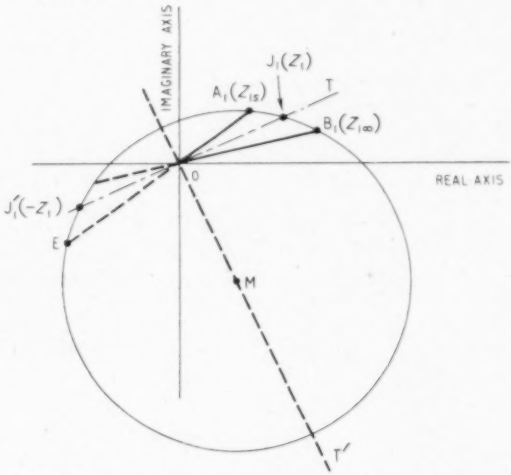

Fig. 32. *Input-impedance diagram. Determination of image impedance* (Z_1) *given open-circuit and short-circuit input impedances* ($Z_{1\infty}$, Z_{1s}) *at terminals* 1.

4.3. *Determination of Image Parameters*

4.3.1. *Image impedance*

It is known that the image impedances Z_1 and Z_2 are related to the open-circuit and short-circuit impedances by the following equations

$$Z_{1s} \cdot Z_{1\infty} = Z_1^2 \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (41)$$

$$Z_{2s} \cdot Z_{2\infty} = Z_2^2 \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (42)$$

For a symmetrical network (**41**) and (**42**) reduce to

$$Z_s \cdot Z_\infty = Z_0^2$$

Equation (41), for example, indicates that Z_1 is the geometric mean of the open-circuit and short-circuit impedances, measured at terminals 1 of the network, and it can therefore be determined geometrically as follows.

Let A_1 and B_1 (Fig. 32) represent Z_{1s} and $Z_{1\infty}$, respectively. J_1, the point representing Z_1, is found at the intersection of the bisector OT to angle B_1OA_1, and the circle through A_1 and B_1, the centre M of which is on the perpendicular OT' to OT.

Since J_1 is on the bisector OT, the phase angle of the impedance represented by J_1 is the arithmetic mean of the phase angles of Z_{1s} and $Z_{1\infty}$. Furthermore, if E and J'_1 are the second intersections of the straight lines OA_1 and OJ_1, respectively, with the circle of centre M, it is known that

$$|OE| = |OB_1|,\ |OJ_1| = |OJ'_1|$$

and

$$|OJ_1| \,.\, |OJ'_1| = |OA_1| \,.\, |OE|.$$

Hence the magnitude of OJ_1 is the geometric mean of the magnitudes of Z_{1s} and $Z_{1\infty}$. It follows that the vector OJ_1 represents Z_1.

4.3.2. *Image transfer constant*

We shall base the determination of the image transfer constant e^{θ}, of a four-terminal network on the following formulae

$$\frac{Z_{1in} - Z_1}{Z_{1in} + Z_1} = \frac{Z_{2L} - Z_2}{Z_{2L} + Z_2} e^{-2\theta} \quad .. \quad .. \quad (43)$$

$$\frac{Z_{2in} - Z_2}{Z_{2in} + Z_2} = \frac{Z_{1L} - Z_1}{Z_{1L} + Z_1} e^{-2\theta} \quad .. \quad .. \quad (44)$$

When the network is symmetrical (43) and (44) reduce to

$$\frac{Z_{in} - Z_0}{Z_{in} + Z_0} = \frac{Z_L - Z_0}{Z_L + Z_0} e^{-2\theta} \quad .. \quad .. \quad (45)$$

If Z_{2L} is made equal to zero, or infinity, Z_{1in} becomes Z_{1s} or $Z_{1\infty}$, respectively, and we obtain the following relations

$$\frac{Z_{1s} - Z_1}{Z_{1s} + Z_1} = -e^{-2\theta} \quad .. \quad .. \quad .. \quad (46)$$

$$\frac{Z_{1\infty} - Z_1}{Z_{1\infty} + Z_1} = e^{-2\theta} \quad .. \quad .. \quad .. \quad (47)$$

Similarly, if Z_{1L} is made equal to zero, or infinity, we obtain

$$\frac{Z_{2s} - Z_2}{Z_{2s} + Z_2} = -e^{-2\theta} \quad .. \quad .. \quad .. \quad (48)$$

$$\frac{Z_{2\infty} - Z_2}{Z_{2\infty} + Z_2} = e^{-2\theta} \quad .. \quad .. \quad .. \quad (49)$$

4.3.3. *Image transfer constant in the impedance diagram*

Let us rewrite equation (46) as follows

$$\frac{|Z_1 - Z_{1s}|}{|Z_1 + Z_{1s}|} e^{j(\alpha_1 - \alpha_2)} = e^{-2\alpha - 2j\beta} \quad .. \quad (50)$$

where α_1, α_2, α and β are defined as below

$$Z_1 - Z_s = |Z_1 - Z_s| \,.\, e^{j\alpha_1} \quad .. \quad .. \quad (51)$$

$$Z_1 + Z_s = |Z_1 + Z_s| \,.\, e^{j\alpha_2} \quad .. \quad .. \quad (52)$$

$$\theta = \alpha + j\beta. \quad .. \quad .. \quad .. \quad (53)$$

If J_1 and J'_1 represent the impedances Z_1 and $-Z_1$ (Fig. 33), then $Z_1 - Z_s$ and $Z_1 + Z_s$ are represented by the vectors A_1J_1 and J'_1A_1, respectively, where A_1 represents Z_{1s}. It follows that -2β is equal to the angle between A_1J_1 and J'_1A_1, while the attenuation ratio, e^{α}, is equal to the square root of the inverse ratio of the above distances. The same relations apply in the case of symmetrical networks provided Z_{1s} and Z_1 are replaced by Z_s and Z_0, respectively.

4.3.4. *Image transfer constant in the Smith Chart*

Let us start from equation (46) and subtract unity from both sides. We obtain

$$1 + e^{-2\theta} = \frac{2Z_1}{Z_1 + Z_{1s}} \quad .. \quad .. \quad .. \quad (54)$$

If now we add unity to both sides we obtain

$$e^{-2\theta} - 1 = -\frac{2Z_{1s}}{Z_1 + Z_{1s}} \quad .. \quad .. \quad (55)$$

The ratio of equations (54) and (55) is

$$\frac{e^{-2\theta} - 1}{e^{-2\theta} + 1} = -\frac{Z_{1s}}{Z_1} = K_{-2\theta} \quad .. \quad .. \quad (56)$$

where $K_{-2\theta}$ is the reflection coefficient of $e^{-2\theta}$.

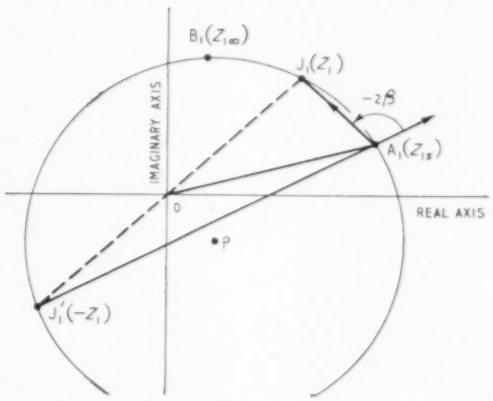

Fig. 33. *Input-impedance diagram. Determination of image transfer constant, given Z_{1s} and $Z_{1\infty}$.*

We can also deduce, in a straightforward manner, the reflection coefficient of $e^{2\theta}$. This is

$$K_{2\theta} = \frac{Z_{1s}}{Z_1} \quad .. \quad .. \quad .. \quad .. \quad (57)$$

If a Smith Chart, with the constant-phase and constant-magnitude circles, is available, it is a simple matter to calculate the reflection co-

efficient of the image loss, given the reflection coefficients of Z_{1s} and Z_1. $K_{2\theta}$, can then, by reference to the above-mentioned circles, in the same chart, be converted into an attenuation and a phase change.

If the reflection coefficient K_1, of the image impedance Z_1, is not given, the value of $K_{2\theta}$ can be determined as follows

From equation (57), we find after squaring

$$(K_{2\theta})^2 = Z_{1s}^2/Z_1^2 \quad .. \quad .. \quad .. \quad (58)$$

Since $Z_1^2 = Z_{1s}.Z_{1\infty}$, we obtain after substitution in (58)

$$(K_{2\theta})^2 = Z_{1s}/Z_{1\infty} \quad .. \quad .. \quad .. \quad (59)$$

Again, using the constant-phase and constant-magnitude circles in the Smith Chart, $K_{2\theta}$ can be determined from the reflection coefficients of Z_{1s} and $Z_{1\infty}$.

If constant-phase and constant-magnitude circles are not already drawn, then the values of Z_{1s} and $Z_{1\infty}$, for use in equation (59) can be determined by the relation

$$Z = \frac{1+K}{1-K} \quad .. \quad .. \quad .. \quad .. \quad (60)$$

which can be easily derived from equation (1).

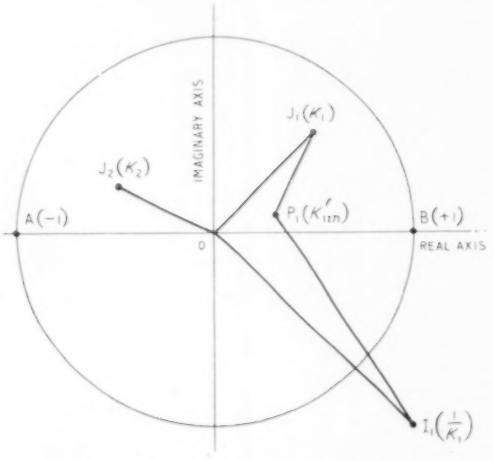

Fig. 34. *Input reflection-coefficient diagram. Determination of image transfer constant given Z_1, Z_2 and K'_{1in} (the input reflection coefficient when the load impedance is equal to unity).*

If A, B and L (Fig. 34), are the points in the Smith Chart representing $K = -1$, $K = +1$ and the reflection coefficient of an impedance Z, then according to (60) the magnitude of Z is given by the ratio of the distances AL to LB, while its phase angle is equal to the angle through which LB must rotate, around L, to coincide with the direction AL.

4.3.5. *Image transfer constant in the Smith Chart. Alternative method*

Let us start from equation (43) and replace all impedances by their values in terms of reflection coefficients. We then obtain

$$\frac{\frac{1+K_{1in}}{1-K_{1in}} - \frac{1+K_1}{1-K_1}}{\frac{1+K_{1in}}{1-K_{1in}} + \frac{1+K_1}{1-K_1}} = \frac{\frac{1+K_{2L}}{1-K_{2L}} - \frac{1+K_2}{1-K_2}}{\frac{1+K_{2L}}{1-K_{2L}} + \frac{1+K_2}{1-K_2}} e^{-2\theta} \quad .. \quad .. \quad .. \quad (61)$$

After cross-multiplying and simplifying we find

$$\frac{K_{1in} - K_1}{1 - K_1 K_{1in}} = \frac{K_{2L} - K_2}{1 - K_2 K_{2L}} e^{-2\theta} \quad .. \quad (62)$$

This can be transformed into

$$\frac{K_{1in} - K_1}{K_{1in} - 1/K_1} = \frac{K_{2L} - K_2}{K_{2L} - 1/K_2} \cdot \frac{K_1}{K_2} e^{-2\theta} \quad .. \quad (63)$$

For a load impedance of unity, $K_{2L} = 0$ and equation (63) becomes

$$\frac{K'_{1in} - K_1}{K'_{1in} - 1/K_1} = K_1 K_2 e^{-2\theta} \quad .. \quad .. \quad (64)$$

Where K'_{1in} represents in equation (64) the input reflection coefficient for a load impedance of unity.

If J_1, J_2, P_1 and I_1 (Fig. 34), represent the reflection coefficients K_1, K_2, K'_{1in} and $1/K_1$, then the left-hand side of equation (64) represents the ratio of the vectors PJ_1 and PI_1. It follows that the image transfer constant can be simply determined from the figure.

A simpler method still is to make use of the open-circuit or short-circuit reflection coefficients. If, in equation (63), we make $K_{2L} = -1$ (i.e., $Z_{2L}=0$), we obtain

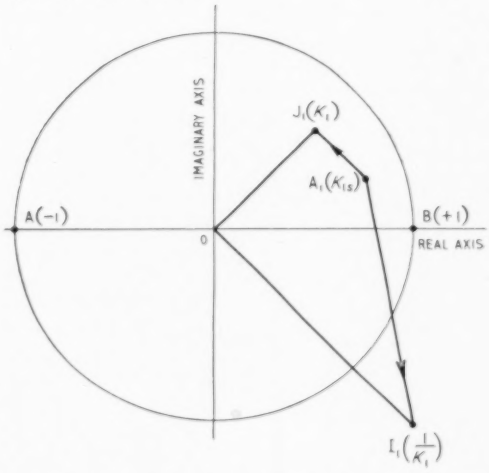

Fig. 35. *Input reflection-coefficient diagram. Determination of image transfer constant given K_1 and K_{1s}.*

$$\frac{K_{1s}-K_1}{K_{1s}-1/K_1}=\frac{1+K_2}{1+1/K_2}\cdot\frac{K_1}{K_2}e^{-2\theta} \quad .. \quad (65)$$

which after simplification reduces to

$$\frac{K_{1s}-K_1}{K_{1s}-1/K_1}=K_1e^{-2\theta} \quad .. \quad .. \quad .. \quad (66)$$

A similar relation can be obtained by making $K_{2L}=-1$ (i.e., $Z_{2L}=\infty$), when equation (63) becomes

$$\frac{K_{1\infty}-K_1}{K_{1\infty}-1/K_1}=-K_1e^{-2\theta} \quad .. \quad .. \quad (67)$$

If J_1, I_1 and A_1 (Fig. 35) represent the reflection coefficients K_1, $1/K_1$ and K_{1s}, then the left-hand side of equation (66) represents the ratio of vectors A_1J_1 to A_1I_1. The image transfer constant can be easily measured off the diagram.

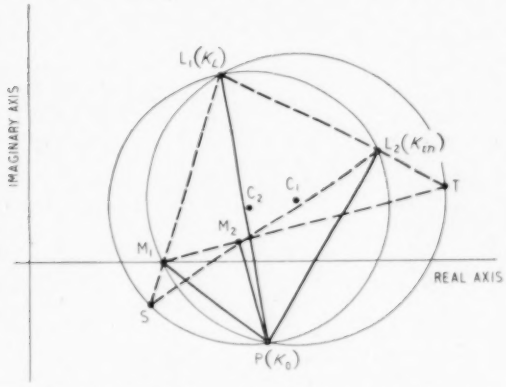

Fig. 36. *Input reflection-coefficient diagram. Determination of reflection coefficient, K_0, of characteristic impedance of symmetrical network, assuming $K_0 \leqslant 0{\cdot}1$.*

In practice it may not always be possible to place I_1 within a reasonable size of chart, especially when K_1 is very small (i.e., when the image impedance Z_1 is nearly equal to unity). In such cases it is possible to use an approximate formula, the accuracy of which increases as K_1 reduces in value.

If we neglect K_1 in the denominator of equation (66) we find

$$K_1-K_{1s}=e^{-2\theta} \quad .. \quad .. \quad .. \quad (68)$$

This indicates that the image transfer constant is given directly by the vector A_1J_1 (Fig. 35), provided K_{1s} is so small as to be negligible compared with $1/K_1$.

4.3.6. *Case of symmetrical networks in the Smith Chart*

Equations (66), (67) and (68) and the corresponding geometrical applications can be used equally well when the network is symmetrical, in which case K_{1s}, $K_{1\infty}$ and K_1 become K_s, K_∞ and K_0, respectively.

Equation (63) becomes

$$\frac{K_{in}-K_0}{K_{in}-1/K_0}=\frac{K_L-K_0}{K_L-1/K_0}e^{-2\theta} \quad .. \quad (69)$$

4.3.7. *Image parameters in the case of nearly-matched low- or moderate-loss networks*

We shall assume that the image impedances of the given network are fairly close to the generator impedance, or in the case of very high frequencies to the characteristic impedance of the measuring line; in other words, that K_1 and K_2 are less than 0·1. This assumption implies that Z_1 and Z_2 are within 10% of unity and we can therefore neglect K_{1in} and K_{2L} in the denominators of equation (63) to obtain

$$\frac{K_{1in}-K_1}{K_{2L}-K_2}=e^{-2\theta} \quad .. \quad .. \quad .. \quad (70)$$

If the given network is known to be symmetrical, then equation (70) reduces to

$$\frac{K_{in}-K_0}{K_L-K_0}=e^{-2\theta} \quad .. \quad .. \quad .. \quad (71)$$

Let L_1 and L_2 (Fig. 36) be the points representing load and input reflection coefficients, respectively, and P the point representing the characteristic impedance of the network. Equation (71) can be written as

$$\frac{\mathbf{PL}_1}{\mathbf{PL}_2}=e^{-2\theta} \quad .. \quad .. \quad .. \quad .. \quad (72)$$

If M_1 and M_2 are two other points having the same relation as L_1 and L_2, then P can be found at the intersection of two circles such as C_1 and C_2 passing through a pair of points out of the four points L_1, L_2, M_1 and M_2 and either of points S or T, where S is the intersection of L_1M_1 and L_2M_2 while T is the intersection of L_1L_2 and M_1M_2.

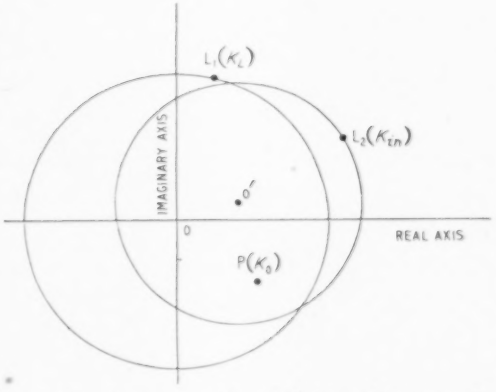

Fig. 37. *Input reflection-coefficient diagram. The constant load reflection circle for symmetrical low-loss network, the characteristic impedance of which is nearly equal to unity.*

This can be proved simply, since according to equation (72) the triangles PM_1M_2 and PL_1L_2 are similar, it follows that the angles M_1PL_1 and M_1TL_1 are equal and consequently the points L_1, M_1, T and P are on a common circle.

Let us now assume that the load reflection coefficient K_L is kept constant in magnitude, while varying in phase. The locus of the point L_1, representing K_L, is a circle of centre O (Fig. 37). The point L_2, representing K_{in}, also describes a circle, the centre of which is O′.

According to equation (72), circle O′ can be derived from circle O, by a rotation around P, followed by a scalar multiplication. It follows that the ratio of the diameters of circle O and O′ is the absolute value of the image loss.

Furthermore, if L_1 and L_2 represent a load and the corresponding input reflection coefficient, then it is possible to determine point P, using the construction indicated at the beginning of this section. Points O and O′ are used instead of points M_1 and M_2. It is worth noting that points L_1 and L_2 need not be on the constant-coefficient circles O and O′. They can be any two points representing a load and the corresponding input reflection coefficient.

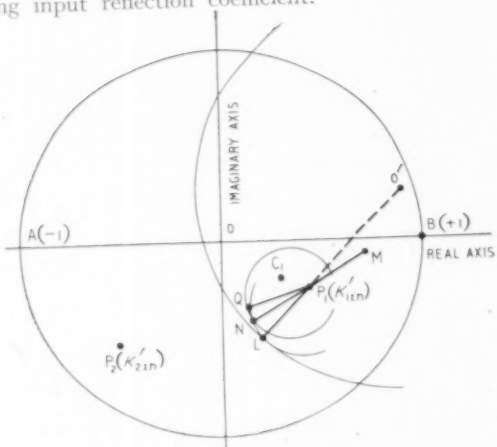

Fig. 38. *Input reflection-coefficient diagram. Determination of insertion loss, given K'_{1in} and K'_{2in}, the primary and secondary reflection coefficients with unity load impedance, and reactance circle C_1, from terminals* 1.

The centre O′ of the input reflection-coefficient circle is the input reflection coefficient due to a load impedance of unity. It happens to be the centre of the constant reflection-coefficient circle because of the assumption that the characteristic impedance of the network is nearly equal to that of the measuring line.

4.4. *Determination of Insertion Loss*[5,6]

The determination of insertion loss when the given network is inserted in an otherwise matched transmission system (i.e., a transmission system in which generator and load impedances are purely real, equal to each other and to the characteristic impedance of a transmission line, if any, connecting the load to the generator) will now be considered.

In general, both terminal pairs are accessible for measurements and in such cases the insertion-loss determination is quite straightforward. In a few cases, however, only one terminal pair is accessible for measurement, and it is then necessary to adopt a slightly modified procedure.

The necessary constructions for symmetrical networks are also treated separately, since they are much simpler than in the case of unsymmetrical networks.

4.4.1. *Insertion loss when both terminal pairs are accessible*

Let P_1, P_2 and C_1 (Fig. 38) be the points representing K'_{1in}, K'_{2in} and the centre of the reactance circle, taken from terminals 1, where K'_{1in} and K'_{2in} are the input reflection coefficients when the network is terminated with unity impedance.

We have seen, in Sect. 3.2, that the reactance

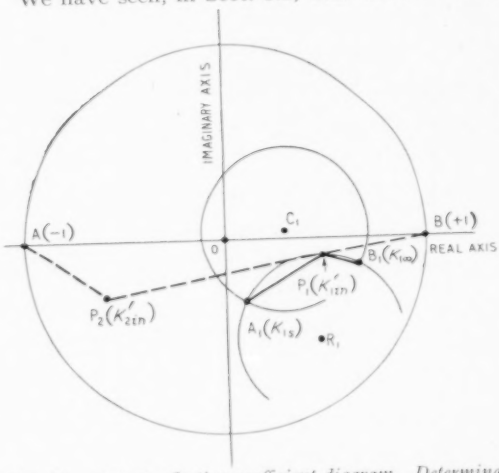

Fig. 39. *Input reflection-coefficient diagram. Determination of insertion loss, given K'_{1in} and K_{1o}, the primary reflection coefficients with the secondary loaded with unity and zero impedance, respectively, and the reactance circle from terminals* 1.

circle C_1 is the inverse of the displaced unity circle of centre O′, the inversion ratio being equal to the complex insertion loss.

It follows that if P_1N is the insertion-loss vector, Q the intersection of P_1C_1 with the reactance circle and L the intersection of P_1O' with the displaced unity circle, then P_1N is the geometric mean of vectors P_1Q and P_1L.

The construction of P_1N, given P_1, P_2 and the

reactance circle, is now obvious and it consists of the following steps. Find O′ such that $\mathbf{OO'} = \mathbf{P_2P_1}$, draw P_1C_1Q and $P_1O'L$ and determine the geometric mean of $\mathbf{P_1Q}$ and $\mathbf{P_1L}$. This is found on the bisector P_1N of angle QP_1L and on the circle through Q and L with its centre on the perpendicular at P_1 to P_1N.

4.4.2. *Insertion loss when only one terminal pair is accessible for measurements*

Let A_1 and P_1 (Fig. 39) be the points representing the input reflection coefficients with loads equal to a short circuit and to unity, respectively, and C_1 the centre of the reactance circle. The insertion-loss determination in this case consists in first determining point P_2, representing the input reflection coefficient looking into terminals 2, when terminals 1 are loaded with unity impedance. Once P_2 is found, we can use the method indicated in the above section to determine the insertion loss.

If we apply equation (30) rewritten here for convenience

$$S^2 = (K_{1in} - K'_{1in})(K'_{2L} - K'_{2in}) \quad .. \quad (30)$$

to the case of a secondary load impedance equal to zero, then $K_{2L} = K'_{2L} = -1$ and consequently

$$S^2 = -(K_{1s} - K'_{1in})(1 + K'_{2in}) \quad .. \quad (73)$$

Similarly, if the load impedance is infinite, $K_{2L} = K'_{2L} = 1$, and we can write

$$S^2 = (K_{1\infty} - K'_{1in})(1 - K'_{2in}) \quad .. \quad (74)$$

From (73) and (74) we deduce

$$1 = -\frac{K_{1s} - K'_{1in}}{K_{1\infty} - K'_{1in}} \cdot \frac{1 + K'_{2in}}{1 - K'_{2in}} \quad .. \quad (75)$$

and, finally,

$$\frac{K_{1s} - K'_{1in}}{K_{1\infty} - K'_{1in}} = -\frac{1 - K'_{2in}}{1 + K'_{2in}} \quad .. \quad .. \quad (76)$$

Equation (76) can be rewritten as

$$\frac{\mathbf{P_1A_1}}{\mathbf{P_1B_1}} = \frac{\mathbf{P_2B}}{\mathbf{P_2A}} \quad .. \quad .. \quad .. \quad .. \quad (77)$$

where B_1 represents the input reflection coefficient for an open-circuited secondary.

Equation (77) establishes the fact that the triangles $P_1A_1B_1$ and P_2AB are similar. It follows that it is a simple matter, given points P_1, A_1, B_1, A and B, to determine point P_2. Since only P_1, A_1 and C_1 are given, it is first necessary to find B_1. We have already seen that this point is the second intersection of the reactance and the resistance circles, and the latter is known to pass through A_1 and P_1 and to be orthogonal to the reactance circle. Hence B_1 and P_2 can be found and then the insertion loss is derived as in the preceding section.

4.4.3. *Alternative insertion-loss determination*

Let P_1 and P_2 (Fig. 40) be the input reflection coefficients of the network with both ends terminated with unity impedance, and Q_1 the centre of a given constant load-reflection-coefficient circle, while circle L is the locus of the load reflection coefficient. We shall assume that P_1, P_2, the circle of centre Q_1 and the load circle L are given.

Let O′ be the centre of the inverse reflection-coefficient circle of the load impedance. We have seen in Sect. 3.2, that OO′ and P_2P_1 are equal vectors. We can now draw a circle of centre O′, the radius r of which is equal to the reciprocal of the magnitude of the load reflection coefficient; i.e., $r = 1/|K_{2L}|$.

If T is the intersection of P_1Q_1 with the Q_1 circle, while D is the intersection of P_1O' with the O′ circle, then, according to equation (33), S, the insertion-loss vector is the geometric mean of vectors P_1T and P_1D.

4.4.4. *Approximate determination of insertion loss for moderate or low-loss networks*

If in equation (39) we neglect the term $|K'_{2in}|^2 |K_{2L}|^2$, on the assumption that K'_{2in} or K_{2L} is small, then the radius of a constant-load reflection-coefficient circle can be expressed by the approximate formula

$$R = |S^2|\,|K_{2L}| \quad .. \quad .. \quad .. \quad .. \quad (78)$$

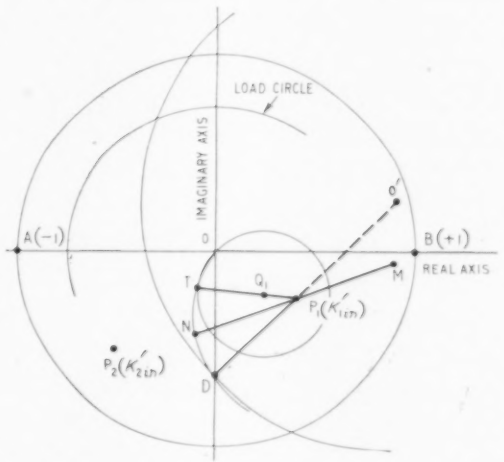

Fig. 40. *Input reflection-coefficient diagram. Determination of insertion loss, given K'_{1in} and K'_{2in} (the primary and secondary reflection coefficients for unity load impedance) and constant reflection-coefficient circle of centre Q_1.*

When K'_{2in} is small, or in other words, if the network input impedance at its secondary terminals, with its primary terminals connected to unity impedance, is nearly equal to the generator

impedance, then the radius of the reactance circle (for which $|K_{2L}| = 1$) is equal to $|S^2|$, the power loss of the network [i.e., the ratio of the power delivered to the load when the latter is connected to the generator (a) via the network and (b) direct].

If K'_{2in} is either not small or unknown, we can determine an input reflection-coefficient circle for a small value of load reflection coefficient, K_{2L}. The radius R of this circle is given by equation (78). It is clear that if the insertion loss is high, then $|S|$ is correspondingly small, and the radius R is very small, since K_{2L} has already been assumed to be small also. It follows that this approximation is not of great practical use if the insertion loss is high. When, however, the insertion loss is small or moderate, the approximation can be used.

Provided either K'_{2in} or K_{2L} is as small as 0·1, which is equivalent to a standing-wave ratio of 1·2, then the insertion loss is known to an accuracy of at least 1%, which is sufficient for most purposes. The final choice of K_{2L} depends mainly upon whether a circle of reasonable size is obtained.

4.4.5. Insertion-loss determination in the case of symmetrical networks

The methods outlined in Sects 4.4.1 to 4.4.4 are equally well applicable to symmetrical networks, with certain simplifications.

In the case of 4.4.1 and 4.4.3, points P_1 and P_2 coincide (we shall call the common point P) and the displaced inverse circles of centre O′ are all concentric with the unity circle.

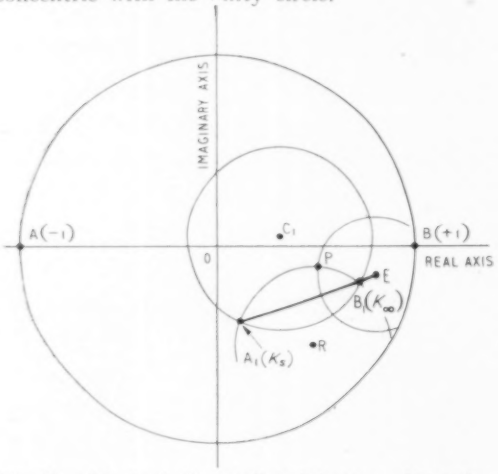

Fig. 41. *Input reflection-coefficient diagram. Determination of point P, representing the input reflection coefficient at primary of symmetrical network, with secondary terminated in unity impedance, given reactance circle C_1 and primary reflection coefficient, K_s, with short-circuited secondary.*

If, however, a well-matched load is not available for the determination of P, the reactance circle and the short-circuit input reflection coefficient can be used instead.

Let A (Fig. 41) be the primary input reflection coefficient with the secondary short-circuited, and C_1, the reactance circle. We first determine B_1, the point representing the primary input reflection coefficient with the secondary open-circuited, using the method indicated in Sect. 4.2.3.

The point P, since it represents the primary reflection coefficient with the secondary terminated in unity impedance, is on the resistance circle. This passes through A_1 and B_1 and it is orthogonal to the reactance circle.

If now we apply equation (20) to symmetrical networks we obtain

$$\frac{K_{in} - K_s}{K_{in} - K_\infty} = -\frac{Z_L}{\dfrac{1 + K_\infty}{1 - K_\infty}} \cdot \frac{1 - K_s}{1 - K_\infty} \quad .. \quad (79)$$

which simplifies to

$$\frac{K_{in} - K_s}{K_{in} - K_\infty} = -\frac{Z_L(1 - K_s)}{1 + K_\infty} \quad .. \quad .. \quad (80)$$

When Z_L is made equal to unity, equation (80) becomes

$$\frac{K'_{in} - K_s}{K'_{in} - K_\infty} = -\frac{1 - K_s}{1 + K_\infty} \quad .. \quad .. \quad (81)$$

The geometrical equivalent of equation (81) is, neglecting the angular information,

$$\frac{|A_1P|}{|B_1P|} = \frac{|A_1B|}{|AB_1|} = \text{constant} \quad .. \quad .. \quad (82)$$

Therefore P is also on a circle centred on A_1B_1 and orthogonal to the reactance circle. The intersection of this circle with the resistance circle determines P.

4.5. *Determination of Transmission Efficiency*[6]

If we define the transmission efficiency of a quadripole as the ratio of output to input power when the quadripole is inserted in an otherwise matched transmission line, we can prove that

$$T_1 = \frac{S^2}{1 - |K'_{1in}|^2} \quad .. \quad .. \quad .. \quad (83)$$

and

$$T_2 = \frac{S^2}{1 - |K'_{2in}|^2} \quad .. \quad .. \quad .. \quad (84)$$

where the transmission efficiency is T_1 when the power flow is from terminals 1 to terminals 2 and T_2 when the direction of power flow is reversed.

If we compare (84) with (39), we see that if the load reflection coefficient, K_{2L}, is made equal to unity (the load is then a pure reactance), then

the radius of the reactance circle taken from terminals 1 is equal to T_2. It follows therefore that in order to determine T_1 we must take the reactance circle from terminals 2.

Since, however, it is not always possible to take measurements from terminals 2, the following method enables the necessary data to be obtained from terminals 1 only. Determine the reactance circle, and the input reflection coefficients first with a matched load, then with a short circuit at the secondary terminals. Using the method indicated in Sect. 4.4.2, determine P_2, the point representing the input reflection coefficient at the secondary, when the primary is terminated in a matched load, and the insertion-loss vector S. Given P_2 and S, determine the secondary reactance circle, by the reverse procedure of Sect. 4.4.1 or obtain T_1 direct from equation (83).

Acknowledgment

The writer wishes to thank the Chief of Research, Marconi's Wireless Telegraph Company Limited, for permission to publish this paper.

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FIELD-STRENGTH CALCULATION

New Method for Mixed Paths

By Kiyohisa Suda

(*NHK, Japan Broadcasting Corporation, Tokyo*)

SUMMARY.—A simple method for the calculation of field strength over mixed paths is explained with the aid of a comparison with measured values of field strength. An equivalent conductivity is obtained from the 'numerical distance' and field strength is calculated easily by conventional curves.

Introduction

THE calculation of medium-wave field strength over a composite path, such as part land and part sea, has been the subject of discussion for many years and it is difficult to estimate field strength in mountainous countries. A number of methods have been tried, however, with various degrees of success, among which three methods, proposed by P. P. Eckersley, Somerville and G. Millington,[1] seem to be practically useful. Values calculated by the three methods were compared with the observed results in the paper by H. L. Kirke,[2] and none of the three methods can give a complete solution for all conditions.

The method proposed by G. Millington, however, was provisionally adopted for the calculation of ground-wave field strength over a mixed path by the 6th General Assembly of the C.C.I.R., Geneva, 1951, and it was recommended that further experimental results should be obtained to verify the method and, in addition, that this problem should be studied theoretically and experimentally.

MS accepted by the Editor, November 1953

A new method (Equivalent Conductivity Method), simpler than the Millington method, has recently been devised by the author and a brief description of the method is given below together with the comparison with measured values of field strength.

Equivalent Conductivity Method

As is well known, A. Sommerfeld has defined the so-called 'numerical distance'[3] for the calculation of ground-wave propagation over a homogeneous plane earth; i.e.,

$$\rho = \left| \frac{k_1^4}{k_2^4} \cdot \frac{k_1^2 - k_2^2}{k_1^2} \cdot \frac{k_1 d}{2} \right| \qquad (1)$$

where
ρ = numerical distance
$k_1 = \omega/c$ (in the air)
$k_2^2 = (\epsilon\omega^2 + j\sigma\omega)/c^2$ (in the earth)
ω = angular velocity
c = light velocity
σ = conductivity
ϵ = dielectric constant
d = actual distance

In the practical case ($|k_2^2| \gg k_1^2$ and $\sigma\omega \gg \epsilon\omega^2$),

a real numerical distance is derived from equation (1), and is

$$\rho = K\frac{d}{\sigma} \quad .. \quad .. \quad .. \quad .. \quad .. \quad (2)$$

$$(K = \text{const.})$$

for a given frequency.

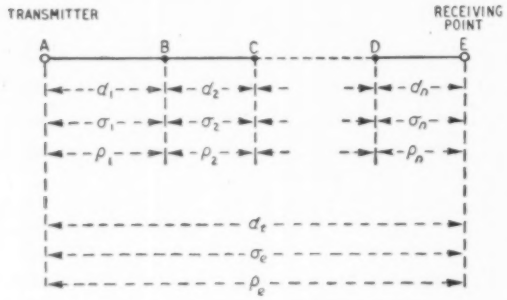

Fig. 1. *Composite path comprising parts A, B, C, D and E; d = actual distance; σ = conductivity and ρ = numerical distance.*

$$= \rho_1 + \rho_2 + \ldots . \rho_n$$

$$= K\left(\frac{d_1}{\sigma_1} + \frac{d_2}{\sigma_2} + \ldots . + \frac{d_n}{\sigma_n}\right)$$

$$= K \times \sum \frac{d_n}{\sigma_n} \quad .. \quad .. \quad .. \quad (3)$$

From equation (3) the equivalent conductivity between A and E is obtained as follows:

$$\frac{d_t}{\sigma_e} = \sum \frac{d_n}{\sigma_n}$$

or $$\sigma_e = d_t \Big/ \sum \frac{d_n}{\sigma_n} \quad .. \quad .. \quad .. \quad (4)$$

Putting σ_e = equivalent conductivity for the path A—E, the field strength at E can easily be obtained from conventional field-strength curves, say, the F.C.C. (Norton) curves.

Experimental Verification

Some verification of the method has been obtained with the aid of experimental measurements carried out in England, Denmark and Japan. The results of measurements in England are quoted from Ref. (2), those in Denmark from Ref. (4), and those in Japan are abstracted from the field intensity measurements of NHK broadcasting stations during 1947–1951. The comparisons are shown in Fig. 2, among which (*c*)—(*f*) refer to the Japanese measurements.

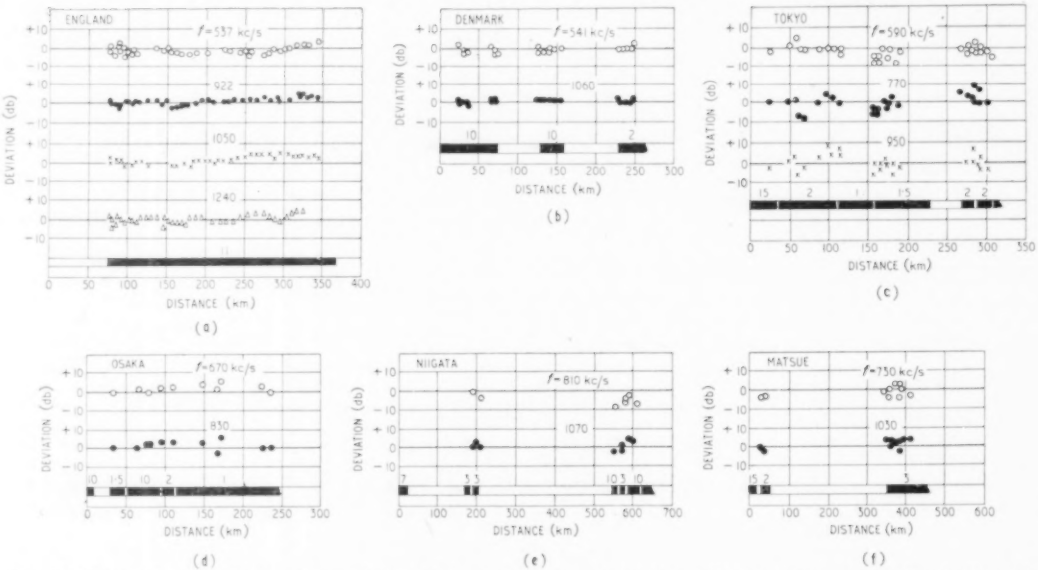

Fig. 2. *The results obtained in various places and for various frequencies f are illustrated here. The points indicate the deviation of the measured from the calculated values. At the bottom, the land parts of the path are shown shaded with their conductivities in* 10^{-14} *c.g.s. e.m.u. marked against them. The other parts represent sea for which the conductivity is taken as* $5{,}000 \times 10^{-14}$ *c.g.s. e.m.u.*

In Fig. 1, conductivities of each of the sections AB, BC, CD, DE, are assumed to be uniform (they are equivalent conductivities considering the geographical conditions as well as mountains, rivers, valleys, buildings, etc.), and we get

$$\rho_e = K\frac{d_t}{\sigma_e}$$

The deviation is smaller than ±10 db and, in general, the calculated values well agree with experimental results.

Conclusion

This method is derived from the 'numerical distance' defined by A. Sommerfeld and has an advantage in the ease of calculation of field strength at many points; e.g., in predicting the coverage of broadcasting stations. The fundamental problem with the method is to estimate conductivities for each of the sections in Fig. 1. and these are determined by geographical conditions, considering various values of conductivities derived from many field-strength measurements.

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NEW BOOKS

Electrical Transients

By L. A. Ware and G. R. Town. Pp. 222 + xi. The MacMillan Company (New York), 10 South Audley Street, London, W.1. Price 33s. ($4.75).

The authors are professors of electrical engineering in the State University of Iowa and the Iowa State College respectively and they have written this book for use "in a course in transient circuits at the senior level". They believe, however, that it could be "used with junior electrical engineering students". They state that no background beyond differential equations is required.

The introduction is largely a refresher course in the solution of ordinary differential equations of simple kind in relation to electrical transients. The Laplace transform is introduced in Chapter 2 and simple applications of it are treated in Chapter 3. The remaining chapters are headed:—Compound, switching and transition transients; The RLC circuit; A.C. applied e.m.f.'s.; Miscellaneous applied e.m.f.'s; Repeated and discontinuous e.m.f. functions; and Vacuum-tube circuits. There are three appendixes.

Although in one sense the content of the book agrees with the title, in another it is much more about the Laplace transform. The whole essence of the book is how to solve problems on transients in electrical circuits by means of the Laplace transform. So much is this the case that the reviewer was quite surprised to find, on closing the book, that its title is "Electrical Transients"; he had quite thought it to be "An Introduction to the Laplace Transform"!

As an introduction to the transform the book is quite a good one and it certainly provides one of the easiest paths available to those who wish to learn how to use it. This is because most of the theory and proofs are omitted, as they should be in an elementary book. However, in spite of this, the treatment is not as good as it could be; a simpler one is possible.

It is surprising that the authors should write down their circuit equations in ordinary differential form and then translate them into Laplace notation. One of the major advantages of the Laplace system, which it shares with the Heaviside operational calculus, is that it enables one to write down the circuit equations exactly as for the steady-state a.c. case merely by writing pL in place of $j\omega L$ and $1/pC$ instead of $1/j\omega C$.

There is, it is true, a reference in a footnote to the possibility of substituting s (which the authors rather inconveniently use instead of the usual p) for $j\omega$. The suggestion is, however, that having written an equation in terms of $j\omega$ one can then substitute s for it. This would be a dangerous practice for one would be liable to overlook terms in which the 'j' had disappeared by squaring.

It is strongly implied in this book that the major advantage of the Laplace transform over other methods is the ease with which initial conditions are dealt. The authors ignore the fact that they are equally easily dealt with in the Heaviside system. They also ignore the real difficulty of all methods, which is that of finding the roots of the denominator of the p-equation when it is of higher degree than the second, as it is in all but fairly simple cases. No mention is made of the possibility of obtaining a solution in the form of a power series in t. Such a solution can always be obtained but it is not always useful, of course; sometimes, however, it is very useful, for conditions may be such that a few terms of the series will suffice.

The authors have slipped on p. 208 where they ascribe to the circuit of Fig. 9.18 a property which it does not possess. Their mathematics are right but they draw the wrong conclusion from them; it is a minor slip which is obvious by inspection of the circuit.

W. T. C.

Fields and Waves in Modern Radio (2nd Edition)

By Simon Ramo and John R. Whinnery. Pp. 576+ viii. Chapman & Hall, Ltd., 37 Essex Street, London, W.C.2. Price 70s.

The first edition of this book appeared in 1944. The present one differs mainly in the use of m.k.s. units, extra problems, the addition of new material and the deletion of some of the less useful parts. The new material includes a chapter on microwave networks and discussions on horns, slot aerials, etc., in the chapter on radiation.

The book, which is of American origin, opens with a chapter on oscillation and wave fundamentals. Chapter 2 deals with the equations of stationary electric and magnetic fields, and Chapter 3 with the solutions to static field problems. Maxwell's equations and high-frequency potential concepts are then treated and followed by circuit concepts, skin effect and circuit impedance elements.

In Chapter 7, the propagation and reflection of electromagnetic waves is discussed and followed by Guided electromagnetic waves, Characteristics of common waveguides and transmission lines, Resonant cavities, Microwave networks, and Radiation.

From this brief summary of its content, the book would appear to be little different from many others. The difference lies in the treatment, which is unusually clear. The subject is one which necessarily involves a fairly high standard of mathematics, but the authors have endeavoured to explain things in physical terms as well as by equations. They have tried, in fact, to make the equations mean something. This is a very laudable effort

and they have, in a great measure, succeeded although not entirely.

On the whole, the treatment is one of the best, if not the best, that we have met. It is least satisfactory in connection with radiation, but this is only in comparison with the rest of this book. The treatment of this especially difficult subject is at least as good as that given elsewhere.

W. T. C.

Radio Laboratory Handbook (6th Edition)

By M. G. Scroggie, B.Sc., M.I.E.E. Pp. 436. Published for *Wireless World* by Iliffe & Sons, Ltd., Dorset House, Stamford Street, London, S.E.1. Price 25s.

This new edition has been almost entirely re-written and is greatly enlarged. It covers the fundamental principles of measurement, sources of power and signals, indicators, standards, and composite instruments such as bridges and Q meters. Later chapters treat the technique of measurement and the reduction of errors.

In spite of its title, the scope is by no means confined to radio. Most of the methods and apparatus are applicable to many branches of electronics where radio-like equipment and components are so widely used.

Radio Valve Data (4th Edition)

Compiled by the Staff of *Wireless World*. Pp. 100. Published for *Wireless World* by Iliffe & Sons, Ltd., Dorset House, Stamford Street, London, S.E.1. Price 3s. 6d.

This reference book contains data on the characteristics and operating conditions of over 2,000 types of British and American valves, cathode-ray tubes and transistors. The data is presented in tabular form and includes valve-base connections. The valves are grouped into current, replacement and obsolete types, a list of equivalents is given and there is an index.

Industrial Electronics

By Dr. R. Kretzmann. Pp. 234. Cleaver-Hume Press, Ltd., 31 Wright's Lane, Kensington, London, W.8. Price 25s.

This book is one of the Philips' Technical Library Series, and it is "concerned with industrial electronics and, by accepted convention, this means electronic circuits in applications other than those for purposes of telecommunication". Part I covers the principles of the various types of valve employed and Part II deals with electronic devices—covering relays, counting circuits, timers, rectifiers, lamp dimmers, speed-temperature control, resistance welding, motor control, h.f. induction heating, ultrasonic soldering and r.f. dielectric heating.

The Electronic Musical Instrument Manual (2nd Edition)

By Alan Douglas. Pp. 221+ix. Sir Isaac Pitman & Sons, Ltd., Parker Street, Kingsway, London, W.C.2. Price 30s.

Practical Wireless Circuits (16th Edition)

By F. J. Camm. Pp. 217. George Newnes Ltd., Southampton Street, Strand, London, W.C.2. Price 10s. 6d.

CABMA Register, 1954-55, of British Products and Canadian Distributors

Pp. 780. Published jointly by Kelly's Directories, Ltd., and Iliffe & Sons, Ltd., for the Canadian Association of British Manufacturers and Agencies. Obtainable in the U.K. from Iliffe & Sons, Ltd., Dorset House, Stamford Street, London, S.E.1. Price 42s.; in Canada, from J. Reg. Beattie, British Trade Centre, Royal Bank Buildings, Toronto; price $7.50.

This book is an alphabetical directory of British manufacturers and distributors and gives details of 4,500 British firms which are classified by products in a Buyers' Guide. The details include their distribution arrangements in Canada. There is a French-English Glossary and an identification of products by Proprietary Names and Trade Marks.

JUBILEE OF THE THERMIONIC VALVE

On 16th November 1954 it will be exactly 50 years since Sir Ambrose Fleming applied for the British Patent for his thermionic valve. The Institution of Electrical Engineers is celebrating this by three lectures on the valve to be given by Sir Edward Appleton, Professor G. W. O. Howe and Dr. J. Thomson on this date. The proceedings will be opened by the Lord President of the Council, the Marquess of Salisbury.

There will also be an exhibition of historical apparatus at the Institution.

BRIT.I.R.E. MEETING

29th September. "Computing Circuits in Flight Simulators", by A. E. Cutler, B.Sc., Ph.D., to be held at London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1, at 6.30.

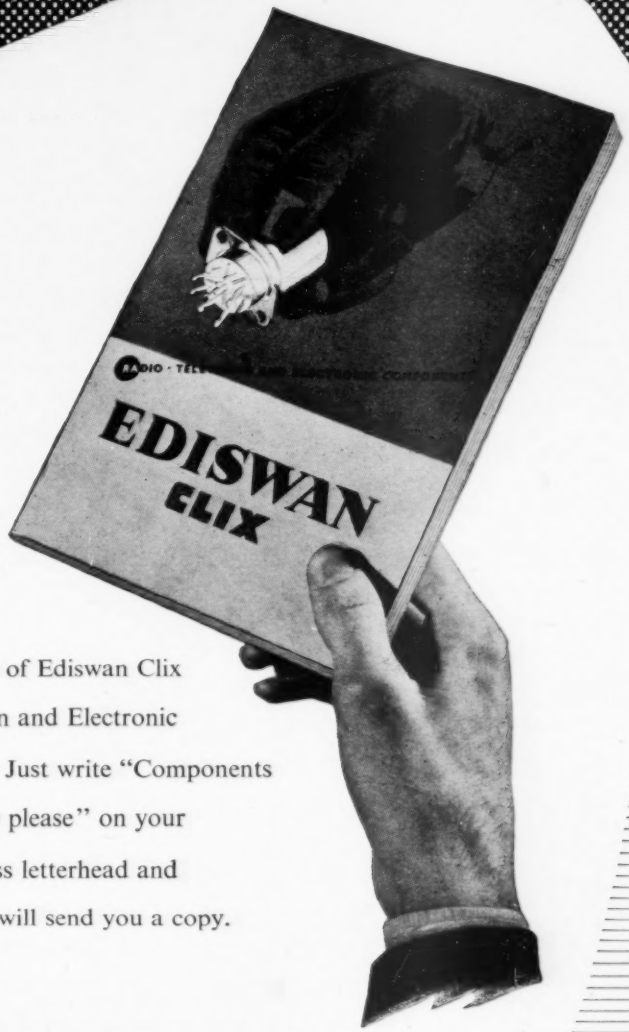
STANDARD-FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

Values for July 1954

Date 1954 July	Frequency deviation from nominal: parts in 10^8		Lead of MSF impulses on GBR 1000 G.M.T. time signal in milliseconds
	MSF 60 kc/s 1429-1530 G.M.T.	Droitwich 200 kc/s 1030 G.M.T.	
1	−0·7	0	+42·0
2	−0·6	0	+40·7
3	−0·6	0	NM
4	−0·6	0	NM
5	−0·6	0	+39·6
6	−0·7	+1	+39·3
7	−0·7	+1	+38·3
8	−0·8	0	+37·7
9	−0·7	+1	+37·2
10	−0·7	0	NM
11	−0·7	0	NM
12	−0·6	0	+36·4
13	−0·7	+1	+36·6
14	−0·6	0	NM
15	−0·6	0	+36·4
16	−0·5	0	+35·9
17	−0·5	0	NM
18	−0·5	+1	NM
19	−0·6	0	+34·8
20	−0·6	+1	+34·6
21	−0·6	+1	+33·5
22	−0·6	0	+32·8
23	−0·5	+1	+32·1
24	−0·5	0	NM
25	−0·5	0	NM
26	−0·5	+1	+27·5
27	−0·5	+1	+29·0
28	−0·5	+2	+27·9
29	−0·5	+1	+26·9
30	−0·5	0	+25·2
31	−0·5	+2	NM

The values are based on astronomical data available on 1st August 1954.
NM = Not Measured.



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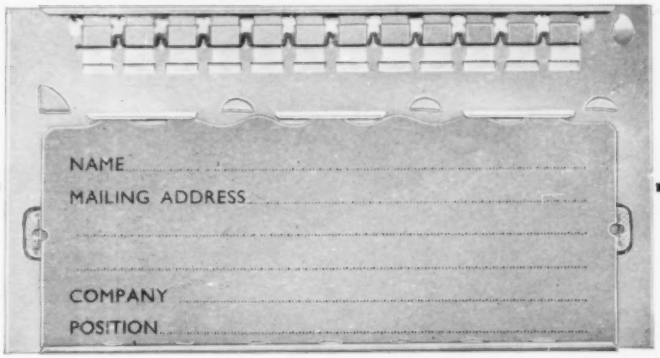

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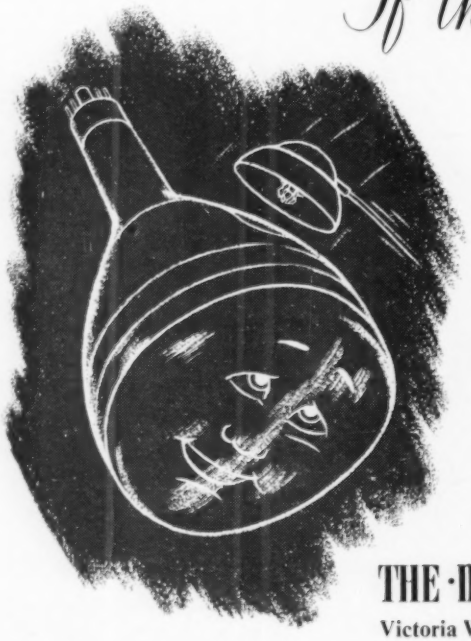

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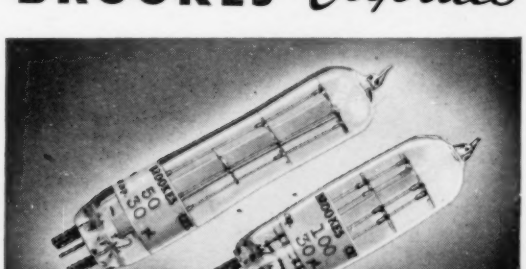

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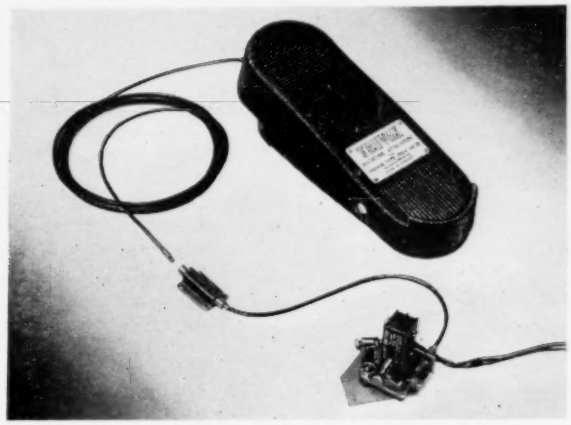

This attachment enables the users of Truvox Tape Decks and Recorders fitted with this Tape Deck to use their machine for dictating. A typist can make transcriptions straight on to the typewriter. List Price 4 gns. Foot pedal only for Recorders previously fitted for this attachment. List Price 3 gns.

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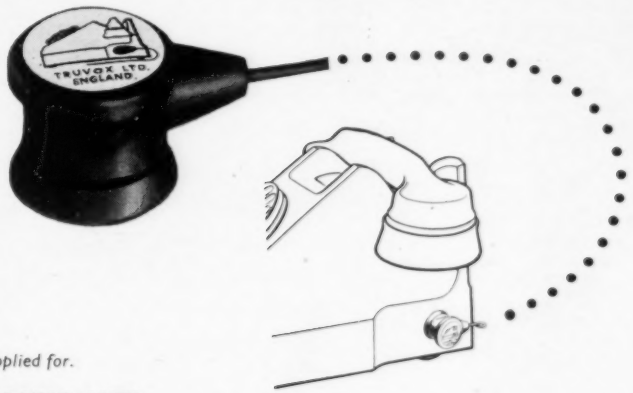

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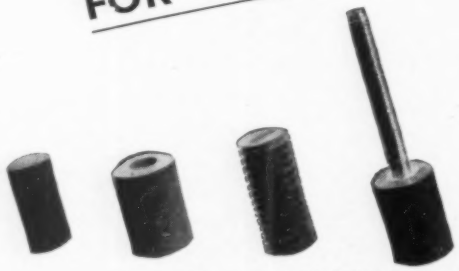

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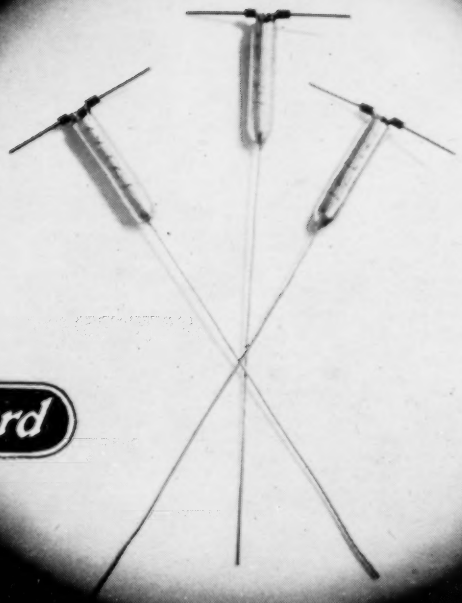

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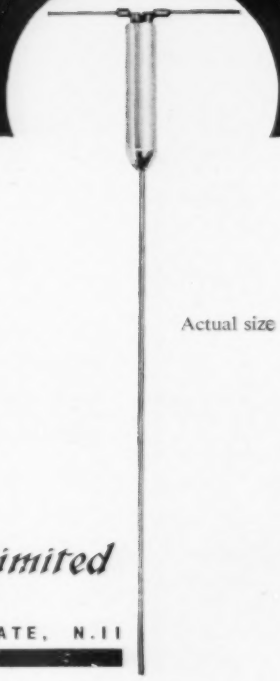

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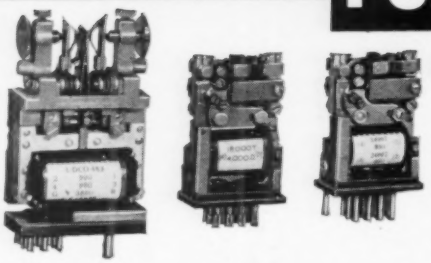

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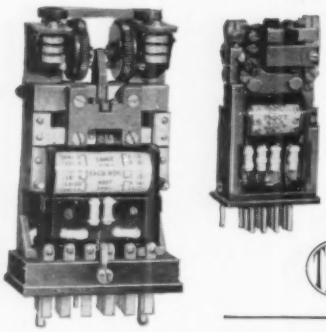

The Carpenter Polarized Relay will respond to weak, ill-defined or short-duration impulses of differing polarity, or it will follow weak alternating current inputs of high frequencies and so provide a continuously operating symmetrical changeover switch between two different sources. Five basic types are available with a wide range of single and multiple windings. Particulars of the type best suited to your purpose will be gladly supplied if you will send us your circuit details.

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The Guided Missiles Division of The ENGLISH ELECTRIC Company Limited, Luton, now wishes to fill a wide variety of vacancies. In particular, the following are required:

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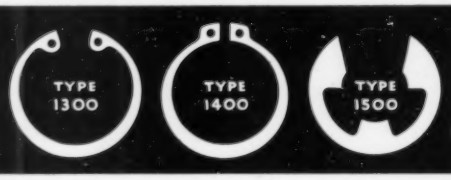

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Fee for the course £2 2s. 0d.

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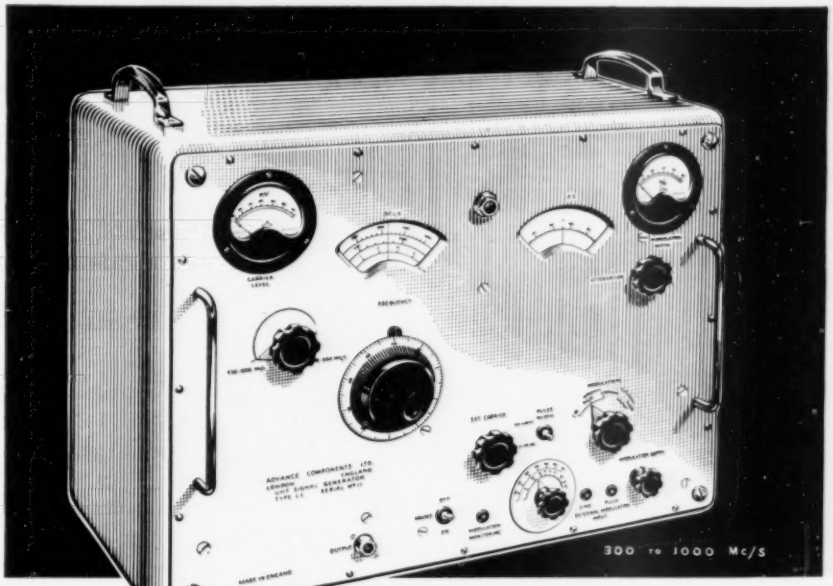

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3. **JUNIOR ENGINEERS,** with a degree or equivalent in physics or engineering. Previous practical experience not essential.

Applications should be made, in strict confidence, with full details, to Box No. 7262 c/o W.E.

E.G.13

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Electronics Engineer with Honours Degree in Physics or Engineering and having experience on the design of electronic computors or simulators.

The vacancy is for someone to take charge of and build up a team which will be working on Simulator Development.

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Good salaries and prospects, pension and life assurance scheme.

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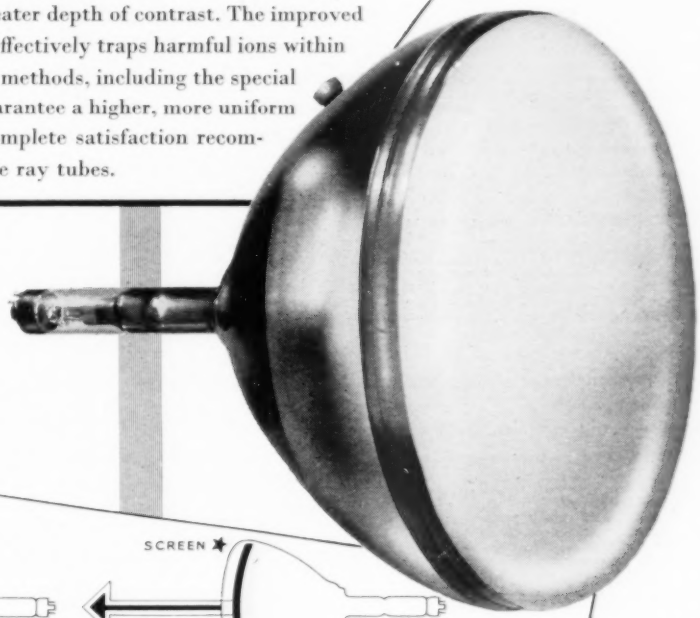

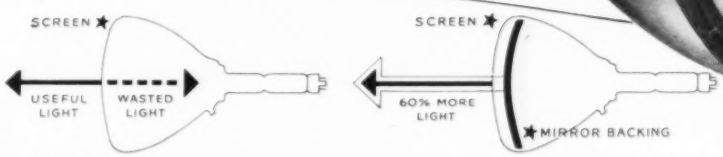

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London salary £480 (at age 23) to £670 (somewhat lower in Provinces). Starting pay up to £580 at age 28 or over on entry. Promotion prospects.

Further particulars and application forms from Civil Service Commission, Scientific Branch, 30 Old Burlington Street, London, W.1, quoting No. S207/54. Application forms must be returned by 8th October, 1954.

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Applicants should preferably have had experience in the Valve or Lamp Industry, or on closely related work. A Science degree, Engineering degree, or a similar qualification would be helpful. An attractive salary is offered in accordance with experience, ability and qualifications.

Please write giving details to the Personnel Officer, Ericsson Telephones, Ltd., Beeston, Nottingham. (Reference LK/3.)

Electronic Digital Computors. Mathematicians with first or second-class Honours Degrees are invited to apply for vacancies in the **English Electric** team at present working in collaboration with the National Physical Laboratories at Teddington in the design and operation of Advanced and Electronic Computors. Please reply to Dept. C.P.S., 336-7 Strand, W.C.2, quoting Ref. 1352.

Electronic Engineer required for the Research Laboratory of **The English Electric Valve Co.** at Chelmsford. Applicants should have a Science Degree or equivalent qualifications, together with considerable experience in the design of circuits for experimental Television and Radar equipment. Apply, giving details of qualifications, experience and salary required to Central Personnel Services, The English Electric Co., Ltd., 336-7 Strand, London, W.C.2, quoting Ref. 440F.

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These appointments are in connection with long-term development of equipment primarily for commercial applications.

Salaries $6,000 upwards according to experience and qualifications. Transportation for family to Canada will also be provided.

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War Office require temporary **Assistant Engineers** for No. 2 War Office Signal Regiment, Droitwich, Worcester and Bampton, Oxon. Salary payable on range £375-£765 (provincial). Starting salary according to age, qualifications and experience. Must be British of British parentage and possess either City and Guilds of London Institute Certificate in Telecommunications Principles (First Class in Grade I or Second Class in Grade II) and first-class certificates in one of following: Telephone Exchange systems, Telegraphy, Radio or Line Transmission/or City and Guilds Intermediate Certificate in Telecommunications Engineering/or first-class certificate in Telecommunication Principles Grade I or above and *two* other certificates in different Telecommunication subjects which should be first class in Grade I or II or first or second class in Grade III or above. Higher National Certificate in Electrical Engineering or Inter B.Sc. will be accepted. Write, giving date of birth, education, full details of qualifications and experience of posts held (including dates) to E.A.394, Appointments Officer, Ministry of Labour and National Service, 1-6 Tavistock Square, W.C.1, by 20th September, 1954. No original testimonials should be sent. Only candidates selected for interview will be advised.

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Ediswan also require Engineers with experience in the fields of electro-medical and industrial electronics design.

Please write to 155 Charing Cross Road, W.C.2, stating experience and salary required. The envelope should be marked "S.P. Lab."

Kelvin & Hughes, Ltd., Barkingside, Essex, have vacancies for Technical Assistants for development of electro-mechanical and electronic instruments (Marine). Work involves initial experiments, sea-going trials, pre-production models and factory liaison. Occasional visits overseas are arranged.

Qualifications—Practical; five years' workshop or drawing office experience. Academic; City and Guilds Telecommunications Final Group Certificate, or equivalent.

Applications stating salary required should be addressed to the Personnel Manager.

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Electronic Digital Computors. Electronics Engineers with an interest in mathematics, are required by **English Electric** to work on fault diagnosis and maintenance of advanced digital computors, in the London area. Please reply to Dept. C.P.S., 336-7 Strand, W.C.2, quoting Ref. 1353.

Birmingham Chartered Patent Agents require the services of a qualified man trained in electronics and dynamics. Excellent prospects. Send particulars of training and career to Box No. 7246.

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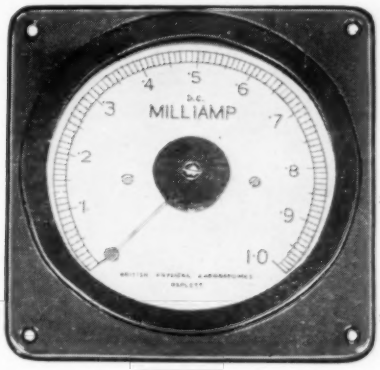

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RANGE FROM 0-500μA

The meter with the dead beat movement

The meter measures only 5″ across and the depth behind the panel is only 2″ yet it retains a scale length of 9″. It is available with spade or knife-edge pointer and special scales can be supplied to customer's specification if required. Send for prices and full details.

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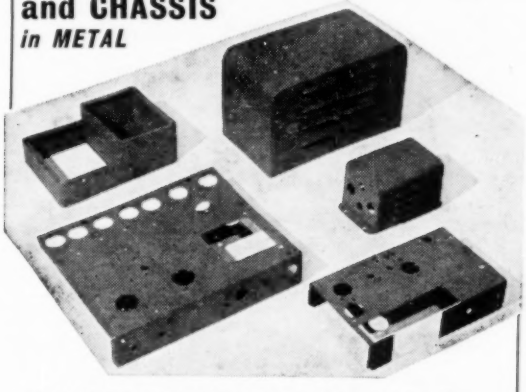

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600 D.C.	—40°C to + 85°C	50pf to 0.001 0.002 to 0.004	A B
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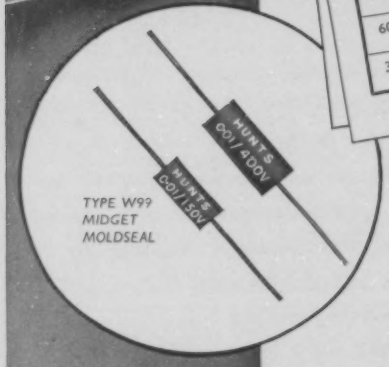

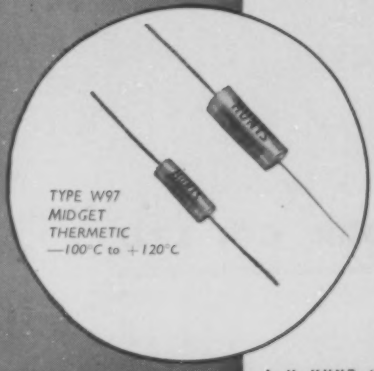

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